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DIGITALLY CONTROLLED "PROGAMMABLE" ACTIVE FILTERS

by

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Digitally Controlled "Programmable" Active Filters

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ABSTRACT

In this research a general purpose digitally controlled analog filter is presented. The novel design is a cascade of second order sections that are individually programmed to achieve any filtering topologies. Two-binary words are used to control the pole frequency ωp and selectivity Qp of each section independently. Each second-order section is a Generalized-Immittance Converter (GIC) biquads which are known for their high stability and low active and passive sensitivity. CMOS switches are used to electronically relocate the minimum number of passive elements to achieve function programmability. Switches are also used to select the number of cascaded sections to realize higher order transfer functions.

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Totally devoted to my father.

I also would like to express my gratitute to Professor Sherif Michael for all his teaching and understanding.

I. INTRODUCTION

A. THE NEED FOR AN ACTIVE PROGRAMMABLE FILTER

The availability of an analog filter with digitally controlled "programmable" coefficients has been the goal of many researchers due to its several attractions. One possibility of a compact, versatile analog filter under remote control opens up many novel and independent application areas. Also, when a programmable filter is combined with a permanent referenced memory which is user-programmable, this would form an economical and versatile device for dedicated stand-alone applications. The need for such a device was motivated by advancement in thick and thin film technologies and continuous upgrading of systems specifications to take advantage of the available technologies to the limits.

Linear analog filtering finds many applications, such as speech processing (recognition or synthesis), geology, instrumentation, communications, process control, adaptive balancing, etc. There has been much emphasis on performing the filtering function digitally, largely because of the ease of varying and optimizing the transfer function. However, for many reasons, such as cost, size, signal processing complexity, and bandwidth, it would be desirable to perform the filter function with linear components yet retain the flexibility of varying the filter parameters digitally.

Recently, the advantages of combining linear components (operational amplifiers, resistors and capacitors) and non linear elements (switches) have been demonstrated using switched capacitor techniques [36-38]. In this research, we are presenting the results of realizing a continuous active device using linear elements and switches controlled by digital signals to achieve a fully programmable filter [32]. Several programming features of the proposed filter are reported. The first feature is the ability of the network to realize the most common filtering functions (function programmability) namely: Low Pass (LP), Band Pass (BP), High Pass (HP), All Pass (AP) and Notch (N) functions, using the minimal set of elements. The second feature is the ability of the network to program (independently) the key parameters of the filtering function chosen (parameter programmability) namely: the pole resonant frequency (ωp) and selectivity (Qp). Finally, the ability to program the network to cascade several sections to achieve higher order filer. All of the above programmability features are performed independently to realize a universal filtering network. In order to demonstrate the idea of this research, it is necessary to introduce some theoretical back ground.

B. LINEAR SYSTEMS

The box at Fig. (1.1) illustrates the concept of the linear system. In the time domain, the system is characterized by its impulse response y(t), which is the output signal y(t) produced in response to an impulse. for an arbitrary input signal u(t), the output signal y(t) is given by the well known convolution integral.

$$y(t) = \int_{-\infty}^{+\infty} g(t-T)\; u(t) \qquad (1.1)$$

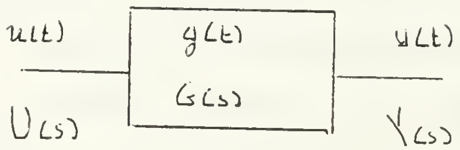


Fig. 1.1 - Black Box Concept of Linear System

The system transfer function G(s) is the Laplace transform of g(t), thus:

$$G(s) = g(t)\; exp(-s\, t) \qquad (1.2)$$

where $s = \sigma + j\omega$

Under the laplace transformation 1.1) becomes:

$$Y(s) = G(s) \cdot U(s) \tag{1.3}$$

so that the transfer function G(s) is the ratio of the output variable to the input variable.

$$G(s) = \frac{Y(s)}{U(s)} \tag{1.4}$$

The most general types of linear systems are consisting of a finite number of lumped, linear and time-invariant elements. The system is characterized by an Nth order ordinary linear differential equation which results, in most general cases, a transfer function G(s) which is a real rational polynomial function of the complex variable S. Thus we can write G(s) in general as

$$G(s) = \frac{P(s)}{E(s)} = \frac{P_M S^M + P_{M-1} S^{M-1} + \cdots + P_0}{\ell_N S^N + \ell_{N-1} S^{N-1} + \cdots + \ell_0} = \frac{\sum_{K=0}^{M} e_K S^K}{\sum_{K=0}^{N} \ell_K \cdot S^K} \tag{1.5}$$

where Pk and ℓk are real numbers so that G(s) is real for real s, and the roots of the polynomial P(s) and E(s) must be real or occur in conjugate pair.

Also by proper multiplications G(s) can take the form

$$G(s) = \frac{P(s)}{E(s)} = C \cdot \frac{\prod_{k=1}^{M} (s - S_{zk})}{\prod_{k=1}^{N} (s - S_{nk})} \tag{1.6}$$

where C is a constant extracted from G(s) such that E(s) and P(s) become monic polynomials (leading coefficients equal to unity), Szk are the transmission zeros and Snk are the natural modes at the system.

C. FILTERS AS A SPECIAL CLASS OF LINEAR SYSTEMS

Linear systems can be distinguished into "SPECTRAL SHAPING NETWORKS" and "FILTERS." The role of filters is one of selecting signals while the role of spectral shaping networks is that of modifying the input signal spectrum in an arbitrary, but predescribed manner. Specifically, we desire that a filter should do as little as possible shaping on signals in its passband; any shaping it is considered a distortion of the signal. On the other hand, networks which perform pulse forming fall within the spectral shaping category.

D. ACTIVE FILTER FUNDAMENTALS

The distinction between passive and active filters is that the first do not require a power source to perform their function while the second do. The motivation behind active

RC filters lies in the desire to have inductorless filter realizations. It is well known that of the three passive R,C&L elements, the inductor is the most nonideal one. This is especially true at low frequencies, where inductors become quite bully and have increased losses or equivalently lower Q-factors.

E. GENERALIZED IMMITTANCE CONVERTOR

One of the methods of active RC filters design consists of simulating the inductances in the LC ladder by active RC networks. This simulation can be based on the principle that we want to find a one port network having an input impedance.

$$Z11 = s*L$$

Various active elements as well as synthesis procedures employing them have been proposed [1-9]. A partial list includes:

(1) Negative Impedance Converter (NIC)

(2) Negative Impedance Inverter (NIV)

(3) Postive Impedance Converter (PIC)

(4) Gyrator

(5) Generalized Impedance Converter (GIC)

(6) Curent Conveyor (CC),and

(7) Operational Amplifier (OA)

Although the introduction of these elements has stimulated research in the area of active network theory, very few elements have made their way to large scale production to become available as off-the-shelf items. The reason for this is mostly an economic one. For a device to become available at low cost, it has to be used in substantially large quantities. It follows that such a device has to be versatile enough to be of use in a number of applications, of which active filter design is only one. The analog circuit design area has found these attributes in the IC operational amplifiers (Op. Amp.)

The IC Op. Amp. is currently the most popular linear active element. It is available from a large number of manufacturers, at reasonable cost and with good performance characteristics. Furthermore, elements engineers have become accustomed to the use of Op. Amp. It is therefore, only natural that the Op. Amp. is becoming the most popular active element in the design of active RC filters, and can be found in NIC's, PIC's, GIC's, and other circuit realizations.

II. THEORETICAL ANALYSIS

A. BIQUADRATIC TRANSFER FUNCTIONS

The filter as a special class of linear system has a transfer function expressed in a polynomial quotient form given as

$$T(s) = \frac{P(s)}{E(s)} = \frac{p_M s^M + p_{M-1} s^{M-1} + \quad + p_0}{g_N s^N + g_{N-1} s^{N-1} + \cdots + g_0} = \frac{\sum_{k=0}^{M} p_k s^k}{\sum_{k=0}^{N} g_k \cdot s^k} \tag{2.1}$$

where Pk and gk are real numbers to that T(s) is real for s, and the roots of the polynomials P(s) and E(s) must either be real or occur in conjugate pairs. Also, in general, the degree of the numerator (deg.[P(s)] = M) is less than or equal to the degree of denominator (deg[E(s)] = N) and the roots of E(s) are in the open-half S-plane. The E(s) is known as the characteristic polynomial or natural mode polynomial of the linear system, and the degree of E(s), that is N, is the order or degree of the system.

A general second-order transfer function or "biquad" function may be written as

$$T(s) = \frac{P_2 s^2 + P_1 s + P_0}{s^2 + g_1 s + g_0} = \frac{P(s)}{E(s)} \qquad (2.2)$$

where P(s) is the loss-pole, or more appropriately here, the transmission-zero polynomial, and E(s) is the natural pole polynomial mode as discussed above. It is a usual practice to express the denominator in terms of _p and Qp, where _p is the natural-mode or resonance frequency and Qp is the natural-mode or quality factor. Thus (2.2) becomes

$$T(s) = \frac{P_2 s^2 + P_1 s + P_0}{s^2 + \frac{\omega_p}{Q_p} + \omega_p^2} \qquad (2.3)$$

The numerator coefficients determine the location of the transmission zeros and hence, the type of filter function the biquad provides. Special cases of interest are:

1. <u>Low Pass (LP)</u>

For which P1=P2=0, thus two transmission zeros are at infinity

$$T(s) = \frac{P_0}{s^2 + \frac{\omega_p}{Q_p} s + \omega_p^2} \qquad (2.4)$$

2. High Pass (HP)

For which p0=p1=0, thus two transmission zeros are at infinity, and

$$T(s) = \frac{P_2 s^2}{s^2 + \frac{\omega_p}{Q_p} s + \omega_p^2} \tag{2.5}$$

3. Band-Pass (BP)

For which Po=P2=0,thus one transmission zero is at infinity while the other is at the origin, and

$$T(s) = \frac{P_1 \cdot s}{s^2 + \frac{\omega_p}{Q_p} s + \omega_p^2} \tag{2.6}$$

4. Notch (N)

For which P1=0 and the two transmission zeros are at $s = \pm j\omega_n$, $\omega_n \underset{HPN}{\overset{LPN}{\gtrless}} \omega_p$ (depending if we have low-pass-notch or high-pass-notch), leading to

$$T(s) = P_2 \cdot \frac{s^2 + \omega_n^2}{s^2 + \frac{\omega_p}{Q_p} s + \omega_p^2} \tag{2.7}$$

5. All Pass (AP)

For which the pair of zeros are at the mirror image location of the pair of poles, that is

$$T(s) = \frac{s^2 - \frac{\omega_p}{Q_p} s + \omega_p^2}{s^2 + \frac{\omega_p}{Q_p} \cdot s + \omega_p^2} \quad P_2 \tag{2.8}$$

B. SENSITIVITY FUNCTIONS

A concern about the design of a filter is how close the resulting response will be to the ideal or desired function. The reason for response deviation from the ideal is the finite tolerances of the RC elements, as well as the nonideal performance of the active elements. In the latter case, not only the gain changes or tolerances have to be considered but the effect of the "limited amplifier bandwidth" on the filter response must also be evaluated. Although effects of initial component tolerances may be "trimmed out" during the initial filter alignments or tuning process, a sensitive design will deviate from the required specifications as time process, due to component variations with temperature, aging, humidity, etc. Note also that a sensitive design might be extremely difficult to tune in the first place, or the initial adjustment will be quite uneconomical.

The answer to tolerance question can be obtained through sensitivity studies. Considerable emphasis has been placed, in the active filter literature, on the study of sensitivity functions and relations. some of the most useful and widely accepted sensitivity functions are the:

1. Magnitude Function Deviation

Assure a filter designed to meet a certain magnitude characteristics T(s) or $T(j\omega)$. One is concerned with the deviation in $|T(j\omega)|$, that is $\Delta|T(j\omega)|$ both in passband and in stopband. Usually it is desirable to express the expected deviation in dB. The deviation $D(\omega)$ dB in the magnitude function may be evaluated as follows. Let the function $|T(j\omega)|$ change to $[|T(j\omega)| + |T(j\ )|]$, then

$$D(\omega) = 20 \log \frac{|T(j\omega) + \Delta T(j\omega)|}{|T(j\omega)|}, (db) \tag{2.9}$$

or

$$D(\omega) = 8.68 \ln \left[1 + \frac{\Delta |T(j\omega)|}{|T(j\omega)|} \right], (db) \tag{2.10}$$

and for small variability (2.10) can be approximated as,

$$D(\omega) \simeq 8.68 \frac{\Delta |T(j\omega)|}{|T(j\omega)|}, (db) \tag{2.11}$$

Thus, the deviation in the magnitude response in nepers is equal to the per unit variability in the magnitude of the transfer function. The problem now reduces to that of evaluating the per unit change in $|T(j\omega)|$. This is not an easy problem since $T(j\omega)$ is a function of many elements with different tolerances and tolerance statistics. Furthermore, the per unit change is function of frequency.

2. Classical Sensitivity

Lets recall the definition of the classical sensitivity, S_x^y where y is a variable of interest, usually a function of many parameters of which x is one, then

$$S_x^y \triangleq \frac{\partial y}{\partial x} \frac{y}{x} = \frac{\partial(\ln y)}{\partial(\ln x)} \tag{2.12}$$

Note that from the above definition, S_x^y is the limit to as Dx -->0. thus, for small variations,

$$S_x^y \simeq \frac{\Delta y/y}{\Delta x/x} \tag{2.13}$$

The usefulness of the classical sensitivity function is evident from (2.13). The per unit or percentage change in y, due to a given per unit or percentage change in x, can be easily obtained by multiplication with S, i.e.,

$$\left(\frac{\Delta y}{y}\right) \simeq S_{x}^{y} \left(\frac{\Delta x}{x}\right) \tag{2.14}$$

3. Gain Sensitivity Product

An important consideration in the evaluation of the sensitivity of a filter parameter as considered in [2] with respect to the closed loop gain is the tolerance on the closed loop due to the open loop gain variability.

Thus, the gain-sensitivity-product, $G.S_{K}^{Y}$ is defined as:

$$G.S_{K}^{Y} \triangleq K \cdot S_{K}^{Y} \tag{2.15}$$

where k is the closed loop gain. We can extend (2.15) to the open loop gain as:

$$G\, S_{A_0}^{Y} \triangleq A_0\, S_{A_0}^{Y} \tag{2.16}$$

and also we can note that:

$$G \cdot S_{K}^{Y} = G \cdot S_{A_o}^{Y} \tag{2.17}$$

and thus then ultimate good is the variability or tolerance rather than the sensitivity, the gain-sensitivity product is a better index for comparing different designs.

4. Determining the Variability of the Transfer Function Amplitude

Assure that the active filter has ℓ resistors, m capacitors and n amplifiers. Let the amplifier k have an open loop gain Aok and, possibly, a closed loop gain Kk. The variability of the magnitude function as given by the Reference [10] is:

$$\frac{\Delta |T(j\omega)|}{|T(j\omega)|} = \sum_{i=1}^{\ell} S_{R_i}^{|T(j\omega)|}\left(\frac{\Delta R_i}{R_i}\right) + \sum_{j=1}^{m} S_{C_j}^{|T(j\omega)|}\left(\frac{\Delta C_j}{C_j}\right) + $$

$$+ \sum_{k=1}^{n} G \cdot S_{K_k}^{|T(j\omega)|}\left(\frac{\Delta A_{ok}}{A_{ok}^{2}}\right) \tag{2.18}$$

Note that each of the sensitivity functions is (2.13) is a function of frequency. In a high order filter realization, the different sensitivity functions might be difficult to evaluate.

C. PROPOSED GIC FILTER ANALYSIS

In order to obtain the transfer functions of the proposed programmable filter [13] shown at Fig. (2.1), nodal analysis was used as follows:

(1) The circuit of Fig. (2.1) was replaced by the one of Fig. (2.2) in which the two operational amplifiers were replaced by the two equivalent dependent voltage sources as can be seen, every element (node, admittance, voltage source, etc.) was labeled and every element was associated to a current direction and a voltage polarity.

(2) The kirchoff current low was written for every node except:

(a) The reference,

(b) any node connected to the reference by a voltage source.

Node	K.C.L.	
1	$i7=-i3$	(2.19)
2	$i2=i5+i6$	(2.20)
3	--	
4	--	
5	$i4=i7+i8$	(2.21)
6	--	
7	--	

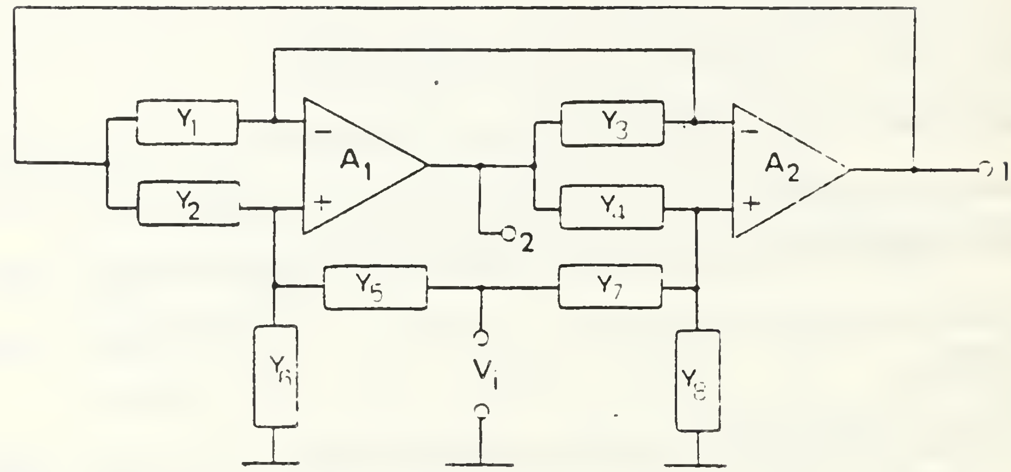


Fig. 2.1 Active Filter Configuration Using the GIC

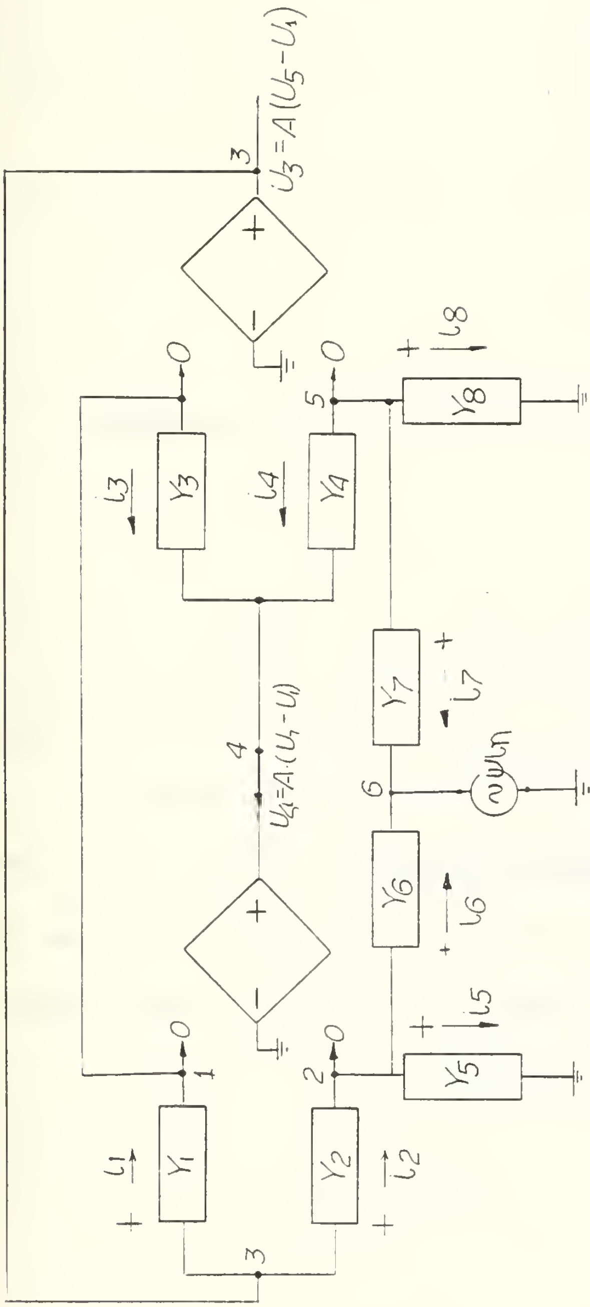


Fig. 2.2 - The Circuit of 2.1 prepared for Nodal Analysis

(3) Every admittance current was expressed in terms of nodal voltages:

$$i1 = Y1[v_3 - v_1] \qquad (2.22)$$

$$i2 = Y2[v_3 - v_2] \qquad (2.23)$$

$$i3 = Y3[v_4 - v_1] \qquad (2.24)$$

$$i4 = Y4[v_4 - v_5] \qquad (2.25)$$

$$i5 = Y5[v_2 - v_6] \qquad (2.26)$$

$$i6 = Y6[v_2] \qquad (2.27)$$

$$i7 = [Y7[v_5 - v_6] \qquad (2.28)$$

$$i8 = Y8[v_5] \qquad (2.29)$$

(4) The source dependencies were listed expressed in terms of nodal voltages,

$$v_4 = A_1[v_2 - v_1] \qquad (2.30)$$

$$v_5 = A_2[v_5 - v_1] \qquad (2.31)$$

and then (substitute) where necessary.

(5) A matrix equation having the unknown voltages V3, V4, and V5 related with the desired transfer functions T1, T2, T3, of the filter was obtained as:

$$\begin{bmatrix} Y_1\left[1+\frac{1}{A_2}\right]+\frac{Y_3}{A_2} & Y_2 & -Y_3-Y_1 \\ Y_2\left[1+\frac{1}{A_1}\right]+\frac{Y_5}{A_2}+\frac{Y_6}{A_1} & -\frac{Y_2}{A_1}-\frac{Y_5}{A_1}-\frac{Y_6}{A_1} & -Y_2-Y_5-Y_6 \\ 0 & Y_4 & -Y_7-Y_8-Y_4 \end{bmatrix} \begin{bmatrix} v_3 \\ v_4 \\ v_5 \end{bmatrix} = \begin{bmatrix} 0 \\ -Y_5 v_{in} \\ Y_7 v_{in} \end{bmatrix} \qquad (2.32)$$

and by substituting $T_1 = \frac{v_3}{v_{in}}$, $T_2 = \frac{v_4}{v_{in}}$ and $T_3 = \frac{v_5}{v_{in}}$ the above matrix equation takes the following form:

$$\begin{bmatrix} Y_1\left[1+\frac{1}{A_1}\right]+\frac{Y_3}{A_2} & Y_3 & -Y_3-Y_1 \\ Y_2\left[1+\frac{1}{A_1}\right]+\frac{Y_5}{A_2}+\frac{Y_6}{A_1} & -\frac{Y_2}{A_1}-\frac{Y_5}{A_1}-\frac{Y_6}{A_1} & -Y_5-Y_6-Y_5 \\ 0 & Y_4 & -Y_7-Y_8-Y_4 \end{bmatrix} \begin{bmatrix} T_1 \\ T_2 \\ T_3 \end{bmatrix} = \begin{bmatrix} 0 \\ -Y_5 \\ -Y_7 \end{bmatrix} \qquad (2.33)$$

So the above matrix equation gives the 3 different responses of the programmable filter as functions of the Y1, . . . Y7 admittances and A1, A2 (Gains) the two Op. Amps. In ideal case [A1=A2=oo] the equation takes the form:

$$\begin{bmatrix} Y_1 & Y_3 & -Y_3-Y_L \\ Y_2 & 0 & -Y_2-Y_5-Y_6 \\ 0 & Y_4 & -Y_4-Y_7-Y_9 \end{bmatrix} \begin{bmatrix} T_1 \\ T_2 \\ T_3 \end{bmatrix} = \begin{bmatrix} 0 \\ -Y_5 \\ -Y_7 \end{bmatrix} \qquad (2.34)$$

which expressed the different realizations discussed by [31].

Now we can express the above matrix equation for the nonideal case in which A1 and A2 are finite and frequency dependent $A_1 = W_1/s$, $A_2 = W_2/s$ where W1 and W2 are the Gain Bandwidth Products (GBWP) of the Op. Amps. A1, A2, respectively.

$$\begin{bmatrix} Y_1\left[1+\frac{s}{W_2}\right]+\frac{Y_3}{W_2}s & Y_3 & -Y_3-Y_1 \\ Y_2\left[1+\frac{s}{W_2}\right]+\frac{s}{W_2}\left[Y_5+Y_6\right] & -\frac{s}{W_1}\left[Y_2+Y_5+Y_6\right] & -Y_2-Y_5-Y_6 \\ 0 & Y_4 & -Y_4-Y_7-Y_8 \end{bmatrix} \cdot \begin{bmatrix} T_1(s) \\ T_2(s) \\ T_3(s) \end{bmatrix} = \begin{bmatrix} 0 \\ -Y_5 \\ -Y_7 \end{bmatrix} \quad (2.35)$$

1. <u>Low Pass (LP) Realization</u>

Can be obtained through T2(s) by substituting Y6=Y7=0, and if we consider the ideal case where (A1=A2=oo), then

$$T_2(s) = \frac{Y_1 Y_5 (Y_4 + Y_8)}{Y_1 Y_4 \cdot Y_5 + Y_2 Y_3 \cdot Y_8} \qquad (2.36)$$

If we further substitute the values of the remaining admittances as proposed in [4], that is Y1=G1, Y2=sC2, Y3=sC3+G3, Y4=G4, Y5=G5, Y8=G8 then (2.36) takes the form

$$T_2(s) = \frac{G_1 G_5 (G_4 + G_8)}{G_1 G_4 \cdot G_5 + G_3 G_8 C_2 s + G_8 C_2 C_3 s^2} \qquad (2.37)$$

which is the form of a low pass transfer function as indicated by (2.4) where

$$P_o = G_1 G_5 (G_4 + G_8) / C_2 C_3 G_8 \tag{2.38}$$

$$\omega_p^2 = \frac{G_1 \cdot G_4 \cdot G_5}{C_2 C_3 G_8} \tag{2.39}$$

and

$$\frac{\omega_p}{Q_p} = \frac{G_3}{C_3} \tag{2.40}$$

2. <u>High Pass (HP) Realization</u>

This one is obtained through T1(s) substituting the following values of the admittances: Y1=G1, Y2=G2, Y3=sC3, Yr=G4, Y5=0, Y6=G6, Y7=sC7, and Y8=G8, and if we assume ideal case (A1=A2-> 00), the

$$T_1(s) = \frac{s^2 C_3 C_7 (G_2 + G_4)}{G_6 G_1 G_4 + C_3 G_2 G_8 \cdot s + G_2 C_3 C_7 s^2} \tag{2.41}$$

expresses the H.P. filter of (2.5) where

$$\omega_p^2 = \frac{G_1 G_4 G_6}{G_2 C_3 C_7} \tag{2.42}$$

$$\frac{\omega_p}{Q_p} = \frac{G_8}{C_7} \tag{2.43}$$

and

$$P_2 = \frac{G_2+G_4}{G_3} \tag{2.44}$$

3. <u>Band Pass (BP) Realization</u>

This is also derived from T1(s) by substituting the following values of admittances: Y1=G1, Y2=G2, Y3=sC3, Y4=G4, Y5=0, Y6=G6, Y7=G7, and Y8=sC8, and if we assume ideal case (A1=A2-->oo), then

$$T_1(s) = \frac{sC_3G_7(G_2+G_6)}{G_1G_4G_6 + sG_2G_7C_2 + C_3C_8G_2s^2} \tag{2.45}$$

which is the form of a BP transfer function as given by (2.5) where

$$\omega_p^2 = \frac{G_1G_4G_6}{G_2C_3C_8} \tag{2.46}$$

$$\frac{\omega_p}{Q_p} = \frac{G_7}{C_8} \tag{2.47}$$

and

$$P_1 = \frac{G_7(G_2+G_6)}{C_8G_2} \tag{2.48}$$

4. <u>Notch (N) Realization</u>

T2(s) expressed the Notch response if the following substitutions have been made. Y1=G1, Y2=G2, Y3=sC3, Y4=G4, Y6=0, Y5=G5, Y7=sC7, Y8=G8, and again assuming ideal case (A1=A2-->oo),

$$T2(s) = \frac{s^2 C_3 C_7 G_2 + G_1 G_5 (G_4 + G_8)}{s^2 C_3 C_7 G_2 + C_3 G_2 G_8 s + G_1 G_4 G_5} \tag{2.49}$$

which is the form of a N transfer function as defined by (2.7) where

$$\omega_n^2 = \frac{G_1 G_5 (G_4 + G_8)}{C_3 C_7 G_2} \tag{2.50}$$

$$\omega_p^2 = \frac{G_1 G_4 G_5}{C_3 C_7 G_2} \tag{2.51}$$

and

$$\frac{\omega_p}{Q_p} = \frac{G_8}{C_7} \tag{2.52}$$

5. <u>All Pass (AP) Realization</u>

This is derived from T1(s) with the following admittances substitutions. Y1=G1, Y2=G2, Y3=sC3, Y4=G4, Y5=G5, Y6=0, Y7=sC7 and Y8=G8, and if once more we assume ideal case, then

$$T_1(s) = \frac{s^2 C_3 C_7 G_2 - s C_3 G_5 G_8 + G_1 G_4 G_5}{s^2 C_3 C_7 G_2 + s C_3 G_2 G_8 + G_5 G_1 G_4} \tag{2.53}$$

which is the response of an All Pass filter as (2.8) indicates where:

$$\frac{\omega_p}{Q_p} = \frac{G_8}{C_7} \quad \text{(for nonminimum phase)} \tag{2.54}$$

and

$$\omega_p^2 = \frac{G_1 G_4}{C_3 C_7} \quad (\quad \text{-- " --} \quad) \tag{2.55}$$

Table 2.1 shows all the realizations proposed by [31]. In our research for designing a programmable filter the No. 1, 3, 7, 9, and 12 realizations were used since they offer the minimum admittance elements change to shift from one to another.

D. SENSITIVITY ANALYSIS

Consider the CGIC circuit shown in Figure (2.2(a)) [12]. Assuming ideal Op. Amps., the chain matrix of the CGIC can be obtained as

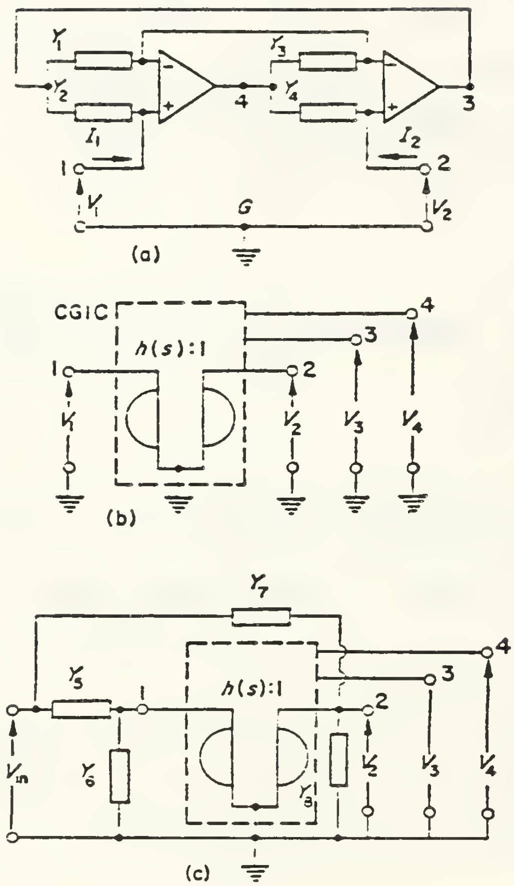


Fig. 2.3 (a) The CGIC Implimentation Using Op. Amps.
(b) Symbolic Representation of the CGIC with created ports 3G and 4G
(c) The Basic Confirguration.

Elements identification for realizing the most commonly used transfer function

Cir. No	$h(s)$	Y_1	Y_2	Y_3	Y_4	Y_5	Y_6	Y_7	Y_8	Transfer function
(1) LP	$\frac{sC_2(sC_3+G_3)}{G_1G_4}$	G_1	sC_2	sC_3+G_3	G_4	G_5	0	0	G_8	$T_2 = \frac{G_1G_5(G_4+G_8)}{G_1G_5G_4+sC_2G_5G_8+s^2C_2C_3G_8}$
(2) LP	$\frac{G_3G_2}{sC_1(sC_4+G_4)}$	sC_1	C_2	G_3	sC_4	0	G_6	G_7	$sC_8^*+G_8^*$	$T_5 = \frac{G_2G_3G_7(1+G_6/G_2)}{G_2G_3(G_7+G_8)+s(C_1G_6G_4+C_8G_7G_3)+s^2C_1C_4G_6}$
(3) HP	$\frac{sC_3G_2}{G_1G_4}$	G_1	G_2	sC_3	G_4	0	G_6	sC_7	$sC_8^*+G_8$	$T_1 = \frac{s^2C_3C_7(G_2+G_6)}{G_1G_4G_6+sC_3G_2G_8+s^2(C_7+C_8)G_2C_3}$
(4) HP	$\frac{sC_3G_2}{G_1G_4}$	G_1	G_2	sC_3	G_4	0	∞	$\frac{sC_7}{1+sC_7R_7}$	0	$T_1 = \frac{s^2C_3C_7(G_2+G_6)}{G_6G_1G_4+sC_3G_1G_4G_6R_7+s^2C_3C_7G_2}$
(5) BP	$\frac{s^2C_2C_3}{G_1G_4}$	G_1	G_2	sC_3	G_4	sC_5	G_6	0	G_8	$T_2 = \frac{sC_5(1+G_8/G_4)G_1G_4}{G_1G_4G_6+sC_5G_1G_4+s^2C_2C_3G_8}$
(6) BP	$\frac{G_2G_3}{sC_1(sC_4+G_4)}$	sC_1	G_2	G_3	$sC_4+G_4^*$	0	G_6	sC_7	G_8	$T_1 = \frac{sC_7G_2G_3(1+G_6/G_2)}{G_2G_3G_8+s(C_7C_2G_3+C_1G_6G_4)+s^2C_1C_7G_6}$
(7) BP	$\frac{sC_3G_2}{G_1G_4}$	G_1	G_2	sC_3	G_4	0	G_6	G_7	$sC_8+G_8^*$	$T_4 = \frac{sC_3G_7(G_2+G_6)}{G_1G_4G_6+sC_3G_2(G_7+G_8)+s^2C_3C_8G_2}$
(8) N	$\frac{G_2G_3}{s^2C_1C_4}$	sC_1	G_2	G_3	sC_4	G_5	0	G_7	sC_8^*	$T_2 = \frac{s^2C_1G_5(C_4+C_8)+G_2G_3G_7}{s^2C_1G_5C_4+sC_8G_2G_3+G_2G_3G_4}$
(9) N	$\frac{sG_2C_3}{G_1G_4}$	G_1	G_2	sC_3	G_4	G_5	0	sC_7	G_8	$T_2 = \frac{s^2C_3C_7G_2+G_1G_5(G_4+G_8)}{s^2C_3C_7G_2+sC_3G_2G_8+G_1G_4G_5}$
(10) N	$\frac{sG_2C_3}{G_1G_4}$	G_1	G_2	sC_3	G_5	G_5	G_6^*	sC_7	sC_7	$T_3 = \frac{(s^2+[(G_1G_4+G_5)/(C_2C_7G_2)])(C_7/(C_7+C_8))}{s^2+s[G_8/(C_7+C_8)]+[G_5G_4(G_5+G_6)]/[C_3G_2(C_7+C_8)]}$
(11) AP	$\frac{G_2G_3}{s^2C_1C_4}$	sC_1	G_2	G_3	sC_4	G_5	G_6^*	G_7	sC_8	$T = \frac{s^2C_1C_4-sG_5G_3C_8+G_3G_7(G_2+G_6)}{s^2C_1C_4(G_5+G_6)+sC_8G_2G_3+G_2G_3G_7}$
(12) ~~KR~~ AP	$\frac{sG_2C_3}{G_1G_4}$	G_1	G_2	sC_3	G_4	G_5	G_6	sC_7	G_8	$T_4 = \frac{s^2C_3C_7(G_2+G_6)-sC_3G_5G_8+G_1G_4G_5}{s^2C_3C_7G_2+sC_3G_1G_8+(G_5+G_6)G_1G_4}$

* These elements can be set equal to zero

Table 2.1 - Elements identification for Realizing the Most Commonly Used Transfer Fuctions

$$[\alpha] = \begin{bmatrix} 1 & 0 \\ 0 & h(s) \end{bmatrix} \tag{2.56}$$

where h(s), the admittance conversion function, is given by:

$$h(s) = Y_2 Y_3 / Y_1 Y_4 \tag{2.57}$$

Two new ports can be created across terminals 3G and 4G as shown symbolically in Fig. (2.3 (b)).

A synthesis procedure is now described which uses the configuration of Fig. (2.3(c)). The transfer functions between the input and output terminals 2, 3, and 4 are readily obtained as

$$v_3/v_i = T_1 = [Y_5 + h(s)\{Y_7(1 + Y_6/Y_2) - Y_5 Y_8/Y_2\}]/D(s) \tag{2.58}$$

$$v_4/v_2 = T_2 = \{Y_5(1 + Y_8/Y_4) - Y_6 Y_7/Y_4 + h(s) Y_7\}/D(s) \tag{2.59}$$

$$v_2/v_2 = T_3 = \{Y_5 + h(s) Y_7\}/D(s) \tag{2.60}$$

where

$$D(s) = Y_5 + Y_6 + h(s)(Y_7 + Y_8) \tag{2.61}$$

The conversion function h(s) and Y5-Y8 can be selected in many different ways and it is found that any second-order transfer function can be realized [12].

Letting

$$Y_i = sC_i + G_i$$

where i=1, 2, 3, or 4, we have from (2.57)

$$h(s) = (sC_2 + G_2)(sC_3 + G_3) / (sC_1 + G_1)(sC_4 + G_4) \tag{2.62}$$

Clearly by omitting one or more conductances and/or capacitances a number of specific conversion functions can be generated.

Most frequently, filters are designed by using Butterworth, Chebychev, Bessel or elliptic approximations in which the transmission zeros are located at the origin, imaginary axis or at infinity. Consequently, the transfer function can be expressed as a product of a second-order transfer functions of the form

$$T(s) = \frac{a_2 s^2 + a_1 s + a_o}{b_2 s^2 + b_1 s + b_o} \qquad (2.63)$$

where a1=a2=0, ao=a1=0, ao=a2=0 or a1=0 for low pass (LP), high pass (HP), band pass (BP) or notch (N) section, respectively.

The coefficients ai's of T(s) for these sections are all positive. These sections can be realized by choosing h(s) in a simple manner such as k_1s, k_2s, $k_3s^2+k_4s$ or their reciprocals. The different ki's (i=1, 2, 3, or 4) are positive constants. By comparing (2.58-2.61) with (2.63), circuits 1-10 in Table 2.1 can be obtained. Circuits 3, 4 and 7 can be regarded as realizations of simple RLC networks [31].

All pass transfer functions are often needed for delay equalization and these can be realized by using second-order transfer functions of the type give by (2.63) where a2=b2, a1=b1, and ao-bo. Second-order sections of this class can be obtained from circuits 11 and 12 of Table 2.1.

Figures (2.3(a)), (2.3(b)), and Table 2.1 show that with the exception of circuit 10, the response is obtained from the output of an operational amplifier. Owing to the low output resistance of the amplifier, any number of sections can be cascaded without isolating amplifiers.

An important criterion of a realization is its sensitivity to element variations. The pole Q factor and the undamped frequency of oscillation from the transfer function of (2.63) are defined as

$$Q_p = \sqrt{b_0 b_1}/b_1 \quad , \quad \omega_p = \sqrt{b_0/b_2} \qquad (2.64)$$

For a Notch section, the Notch frequency is defined by

$$\omega_n = (a_0/a_2)^{1/2} \qquad (2.65)$$

and the multiplier constant can be taken to be

$$H_N = a_0/b_0 \quad \text{or} \quad a_2/b_2 \qquad (2.66)$$

for $\omega n > \omega p$ or $\omega n < \omega p$, respectively.

Similarly for the LP, HP, and BP sections

$$H_{LP} = a_0/b_0 \, , \quad H_{HP} = a_2/b_2 \quad \text{and} \quad H_{BP} = a_1/b_1 \qquad (2.67)$$

For an All Pass Section, let

$$Q_z = \frac{(a_0 a_2)^{1/2}}{a_1}, \quad \omega_z = (a_0/a_2)^{1/2} \text{ and } H_{AP} = a_2/b_2 \tag{2.68}$$

The sensitivity of a quantity x with respect to variations in an element e is given by

$$S_e^x = \frac{e}{x} \frac{\partial x}{\partial e}$$

For ideal amplifiers, the use of (2.64)-(2.68) and Table (2.1) leads to

$$0 \leq |S_e^x| \leq 1 \tag{2.69}$$

where x represents any one of the quantities defined by (2.64)-(2.68) and e represents any capacitance of conductance. In addition, it can be shown that

$$\sum S_e^x = 0$$

and

$$\sum \left| S_e^{Q_p} \right| = \sum \left| S_e^{Q_z} \right| = 4 \qquad (2.70)$$

For amplifiers with a finite open-loop gain A, according to [31] the circuit of (Fig. 1(c)) gives

$$v_k / v_i = N_k(s) / D(s) \qquad (2.71)$$

where k=2, 3, 4, and

$$\left.\begin{aligned} D(s) &= F_1 Y_1 + F_2 Y_3 + (1+F_1)(1+F_2)(Y_1/A_1 + Y_3/A_2 + Y_1/A_1A_2 + Y_3/A_1A_2) \\ F_1 &= (Y_5 + Y_6)/Y_2 \\ F_2 &= (Y_7 + Y_8)/Y_4 \end{aligned}\right\} \qquad (2.72)$$

Consider realizations in which h(s) = k1s, such as the circuits 7(BP), 3(HP), 9, 10(N), where

$Y_1 = G_1$, $Y_2 = G_2$, $Y_3 = sC_3$, $Y_4 = G_4$, $Y_5 = G_5$, $Y_6 = G_6$, $Y_7 = sC_7 + G_7$ and
$Y_8 = sC_8 + G_8$

(2.73)

For real amplifier gains such that

$$A_1 = A_2 = A_0 \quad \text{and} \quad A_0 \gg 1$$

(2.72) gives

$$D(s) = F_1 Y_1 + F_2 Y_3 + (1+F_1)(1+F_2)(Y_1 + Y_3)/A_0 \qquad (2.74)$$

From (2.64), (2.73), and (2.74), the Q-factor and the undamped frequency of oscillation can be obtained as

$$Q_{Pa} = Q_P \left\{ 1 + \frac{X_4}{X_1 A_0} \right\}^{1/2} \left\{ 1 + \frac{X_5 G_4}{(C_7 + C_8) A_0} \right\}^{1/2} \Big/ \left\{ 1 + \frac{C_3 X_4 + G_1 X_5}{C_3 X_2 A_0} \right\},$$

$$\omega_{Pa} = \omega_P \left\{ \left(1 + \frac{X_4}{X_1 A_0}\right) \Big/ \left(1 + \frac{X_5 G_4}{(C_7 + C_8) A_0}\right) \right\}^{1/2} \qquad (2.75)$$

where

$$Q_P = \{ (G_5 + G_6)(C_7 + C_8) G_1 G_4 / (G_7 + G_8)^2 G_2 G_3 \},$$

$$\omega_P = \{ (G_5 + G_6) G_1 G_4 / (C_7 + C_8) G_2 C_3 \}^{1/2},$$

$$X_1 = (G_5 + G_6)/G_2,$$

$$X_2 = (G_7 + G_8)/G_4, \qquad (2.76)$$

$$X_3 = (C_7 + C_8)/G_4,$$

$$X_5 = (1 + X_1)/X_3 \quad \text{and} \quad X_4 = (1 + X_1)(1 + X_2)$$

The sensitivities of Qpa and pa with respect to the amplification Ao can be written as

$$S_{A_0}^{Q_{Pa}} \doteq \frac{-1}{2A_0} \left[\frac{X_4}{X_1} + \frac{X_5}{X_3} \quad \frac{2X_4}{X_2} - \frac{2G_1}{C_3} \frac{X_5}{X_2} \right],$$

$$S_{A_0}^{\omega_{Pa}} = \frac{1}{2A_0} \left[\frac{X_5}{X_3} - \frac{X_4}{X_1} \right] \qquad (2.77)$$

The use of (2.77) and (2.78) leads to

$$S_{A_0}^{Q_{Pa}} = \frac{-1}{2A_0} \left[\left(1 + \frac{1}{X_1}\right)(1 + X_2) + (1 + X_1) \left\{ 1 - 2\left(1 + \frac{1}{X_2}\right) - \frac{2 Q_P^2 X_2}{X_1} \right\} \right] \qquad (2.78)$$

by assuming that

$$\left(1+\frac{1}{X_1}\right) - X_1 \ll 2X_2(1+X_1) + 2Q_p^2 X_2\left(1+\frac{1}{X_1}\right) \qquad (2.79)$$

Eq.(2.78) reduces to

$$S_{A_o}^{Q_{pa}} = \frac{Q_p}{A_o}\left[Q_p X_2\left(1+\frac{1}{X_1}\right) + (1+X_1)/(Q_p X_2)\right] \qquad (2.80)$$

Straight forward differentiation shows that $S_{A_o}^{Q_{pa}}$ is minimum when

$$X_1 = 1 \;,\; X_2 = 1/Q_p \qquad (2.81)$$

From (2.79) and (2.81) the analysis is valid provided that $4Q_p + 4/Q_p \gg 1$ which is clearly satisfied in practice.

From (2.80) and (2.81) the minimum sensitivity to variations in Ao is derived as

$$S_{A_o}^{Q_{pa}} \doteq 4Q_p/A_o \qquad (2.82)$$

The corresponding value of $S_{A_0}^{\omega pa}$ is given by

$$S_{A_0}^{Q pa} \doteq -\frac{1}{A_0 Q_p} \qquad (2.83)$$

LP realizations as circuits 1 and 2 can be obtained by using a conversion function of the form or its reciprocals. The admittances Y1 to Y8 are chosen as

$$Y_1 = G_1,\ Y_2 = sC_2,\ Y_3 = sC_3 + G_3,\ Y_4 = G_4,\ Y_5 = G_5,\ Y_6 = Y_7 = 0,\ Y_8 = G_8 \qquad (2.84)$$

in order to obtain a conversion function of the form $k_3 s^2 + k_4 s$.

Now Qpa and ωpa are obtained as

$$Q_{pa} = \frac{Q_p\{1+(1+G_8/G_4)(1+G_3/G_1)/A_0\}^{1/2}\{1+(1+G_4/G_8)/A_0\}^{1/2}}{1+(1+G_4/G_8)(1+G_1/G_3)/A_0+(1+G_4/G_8)(C_3G_5/G_3C_2A_0)} \qquad (2.85)$$

$$\omega_{pa} = \omega_p\{1+(1+G_8/G_4)(1+G_3/G_1)/A_0\}^{1/2}/\{1+(1+G_4/G_8)/A_0\}^{1/2} \qquad (2.86)$$

$$\left.\begin{aligned} Q_p &= \{(C_3G_1G_4G_5)/(C_2G_3^2G_8)\}^{1/2} \\ \omega_p &= \{(G_1G_4G_5)/(C_2C_3G_8)\}^{1/2} \end{aligned}\right\} \qquad (2.87)$$

The sensitivity of Qpa with respect to variations in Ao can be minimized following the approach used earlier. It is found <u>that for minimum sensitivity</u>[31]

$$G_4 = G_8 \,, \quad G_1 = Q_p G_3 \tag{2.88}$$

The minimum value of $S_{A_o}^{Q_{pa}}$ can be shown to be

$$S_{A_o}^{Q_{pa}} = 4Q / A_o \tag{2.89}$$

and the corresponding value of $S_{A_o}^{\omega_{pa}}$ is given by

$$S_{A_o}^{\omega_{pa}} = -1 / A_o Q_p \tag{2.90}$$

The above sensitivity analysis can be extended to realizations using any other type of conversion function [31].

Equation (2.69) shows that the sensitivities to passive element variations are independent of the selectivity. Furthermore, the sensitivities with respect to variations in the amplifier gain are low. The proposed realizations are seen to have similar sensitivity properties as the low sensitivity realizations reported in [21-27].

E. STABILITY

It has been shown elsewhere [20] that some networks using GIC's can be conditionally stable where a circuit can lock in an unstable mode during activation (just after switching on the power supply). In this section the stability properties of the configuration show in Fig. (2.2(c)) are examined.

The natural frequencies of the circuit in Fig.(2.3(c)) are the zeros of the characteristic polynomial D9s) as given by (2.72). The differential open-loop gain of a frequency compensated Op. Amp., in a bounded frequency range $0 \le \omega \le \omega_m$ is

$$A = A_o . \omega_c / (s + \omega_c)$$

where Ao and wc are the d.c. gain and the cutoff frequency respectively, and $0 \le A_o \le A_{max}$. In the frequency range $\omega << \omega_c$ the amplifier gains A1 and A2 can be assumed to be real. For any second-order transfer function the coefficients of D(s) are seen to remain positive for any attainable pair of A1 and A2. this is due to the absence of negative terms in D(s). Therefore, the zeros of D(s) will remain the left-half s-plane and low frequency unstable modes cannot arise during activation.

III. PROGRAMMABLE GIC FILTER

A. GENERAL

Signal processing devices evolved considerably over the last several years. The progress was motivated by the advancement in film and semiconductor technologies, as well as the continuous upgrading of systems specifications to take advantage of the available technologies to the limits.

Linear filtering finds many applications, such as speech processing (recognition or synthesis), geology, instrumentation, communications, process control, adapting balancing, etc. There has been much emphasis on performing the filter function digitally, largely because of the ease of varying and optimizing the transfer. However, and for many reasons, such as cast size signal processing complexity, and bandwidth, it would be desirable to perform the filter function with linear components, yet retain the flexibility of varying the filter parameters digitally.

Recently, several advantages of combining linear components (amplifiers and capacitors) and nonlinear elements (switches) have been demonstrated using MOS switched capacitor techniques [31, 32]. Here, we are presenting the results of realizing a continuous active device using linear elements and switches controlled by digital signals to achieve fully programmable filters.

Our research addressed two different aspects of programmability namely;

(1) Programming the filter topology using a minimal set of elements to obtain any type of filtering function desired, e.g., LP, HP, BP, N and AP.

(2) Programming the filter's transfer function parameters, (pole resonant frequency ωp and quality factor Qp) for a chosen type of filtering function.

B. THE PROPOSED GIC PROGRAMMABLE FILTER

The basic active network considered as the heart of the GIC programmable filter is the GIC structure [31] of Fig.(3.1), whose superior performance was established in the literature [10,32]. The filter transfer function was derived using loop analysis in Chapter II.

Table (3.1), illustrates that for any of the LP, HP, BP, N and AP realizations, five resistors, two capacitors and two Op. Amps., are required. also, the transfer function of each realization is shown. The passive elements are connected to the different nodes, shown in fig.(3.2), for the different realizations. A set of MOS bilateral switches controlled by a digital binary word, are used to interchange the elements to achieve the different types of filter realization shown in Fig.(3.3). The truth table of the switch control logics is shown in Table (3.2). Fig.(3.3(a)) illustrates the CMOS logical circuit for realizing this truth table. While four of the resistors are equal and of value R each, the fifth

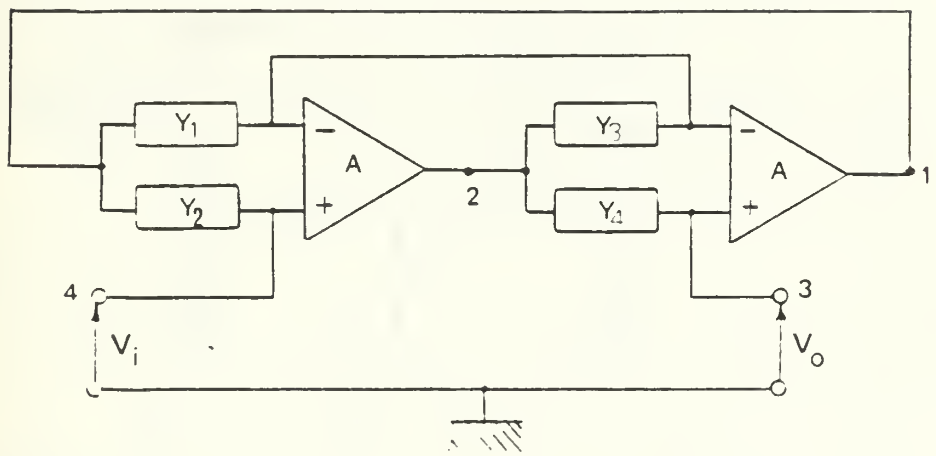


Fig. 3.1 - The Generalized Immittance Converter (GIC) Implementation Using OA's .

Filter Type	Y_1	Y_2	Y_3	Y_4	Y_5	Y_6	Y_7	Y_8	Transfer Function
LP	G	C	$C+\frac{G}{QP}$	G	G	0	0	G	$T_2 = 2Wp^2/D(s)$
HP	G	G	C	G	0	G	C	$\frac{G}{QP}$	$T_1 = 2S^2/D(s)$
BP	G	G	C	G	0	G	$\frac{G}{QP}$	C	$T_1 = 2(Wp/Qp)^s/D(s)$
N	G	G	C	G	G	0	C	$\frac{G}{QP}$	$T_2 = (S^2+W_n^{\ 2})/D(s)$
AL	G	G	C	G	G	0	C	$\frac{G}{QP}$	$T1=(s^2-\frac{wp}{Qp}s+wp^2)/D(s)$

where $T(s) = N(s)/D(s)$ and $D(s) = S^2 + (Wp/Qp)S + Wp^2$

TABLE 3.1. - The Elements Identification for Different Realizations of the GIC Filter .

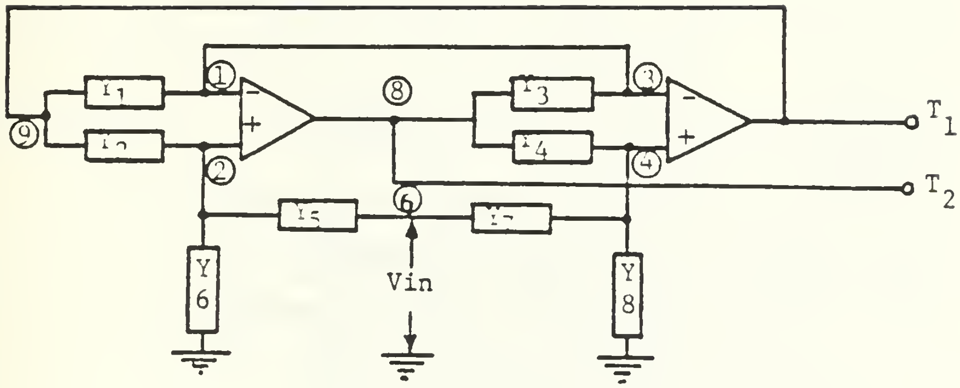


Fig. 3.2 - Schematic Diagram of the Programmable GIC Filter Showing the Controlled Nodes .

To be Connected in Case of	Component Represented Realized	Node in Fig.1	Component & Switches	Node in Fig.1	Component Represented Realized	To be connected in case of
A, L, H, B, N	R_2, R_5	(2)	R, S_1, S_2	(9) (6)	R_2 R_5	H, B L, N, A
N, A L, H, B	R_2 R_6, R_8	(9) GND	S_{15}, S_{16}, R, S_3, S_4	(2) (4)	R_6, R_2 R_8	H, B, N, A L
L H, B, N, A	C_2 C_7, C_8	(2) (4)	S_{13}, S_{14}, C, S_5, S_6, S_7	(9) (6) GND	C_2 C_7 C_8	L H, N, A B
L H, B, N, A	RQ_3 RQ_7, RQ_8	(8) (4)	S_{11}, S_{12}, RQ, S_8, S_9, S_{10}	(3) (6) GND	RQ_3 RQ_7 RQ_8	L B H, N, A
H, B, A L, N	T_1 T_2	(9) (8)	S_{29}, S_{30}	Output		L, H, B, N, A

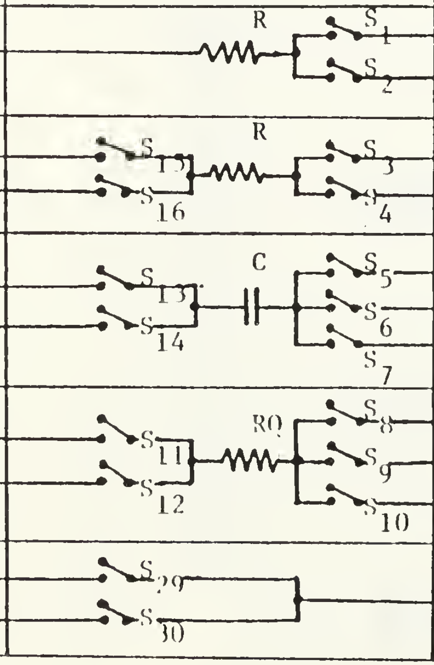


Fig. 3.3 - Different Elements Realizations and the Corresponding Switches Used for Digitally Selecting the Filtering Type .

Binary Input	Filter \ Switch	S_1	S_2	S_3	S_4	S_5	S_6	S_7	S_8	S_9	S_{10}	S_{11}	S_{12}	S_{13}	S_{14}	S_{15}	S_{16}	S_{29}	S_{30}
0 0 0	Low Pass	0	1	0	1	1	0	0	1	0	0	1	0	1	0	0	1	0	1
0 0 1	High Pass	1	0	1	0	0	1	0	0	0	1	0	1	0	1	0	1	1	0
0 1 0	Band Pass	1	0	1	0	0	0	1	0	1	0	0	1	0	1	0	1	1	0
0 1 1	Notch	0	1	1	0	0	1	0	0	0	1	0	1	0	1	1	0	0	1
1 0 0	All Pass	0	1	1	0	0	1	0	0	0	1	0	1	0	1	1	0	1	0

TABLE 3.2 -The Truth Table of the Switches Logic Used to Select the Filtering Function .

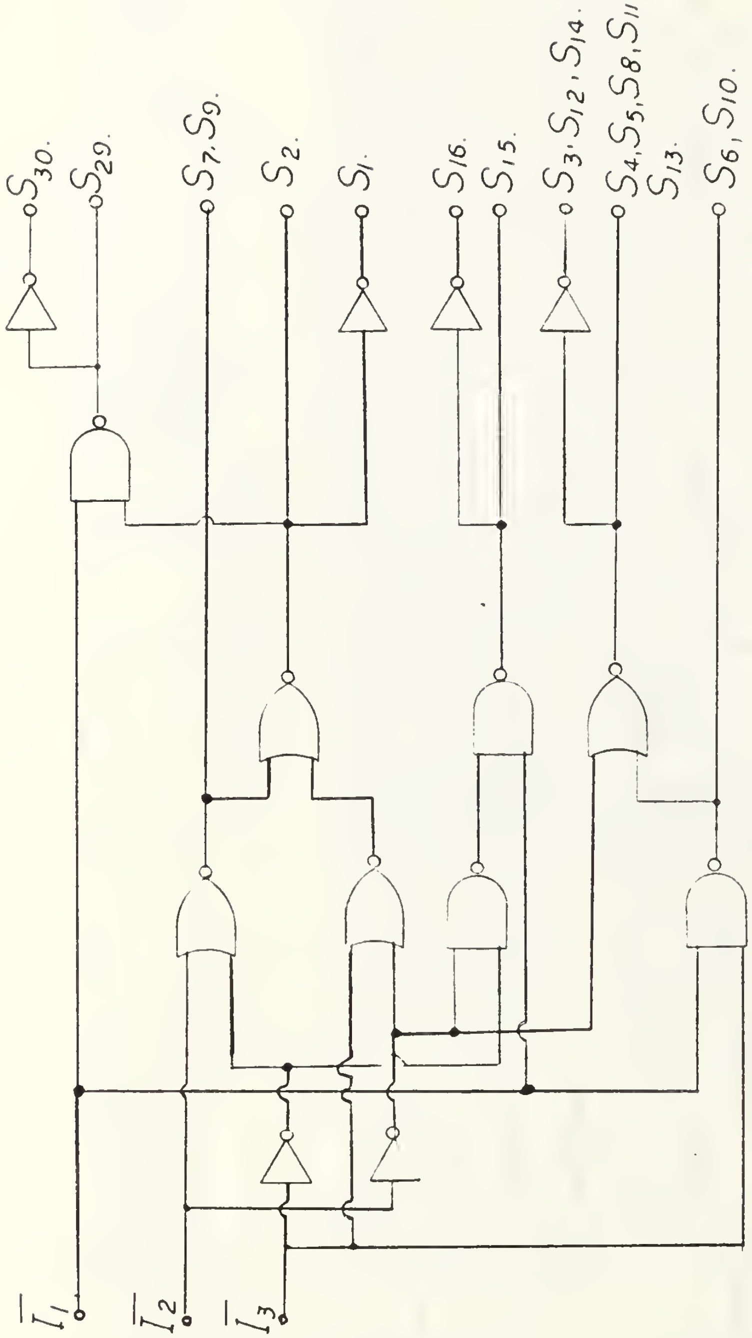


Fig. 3.3a The CMOS Logic Diagram Used to Control the Analog CMOS Switches of Fig. and to Realize the Truth Table .

resistor is the Qp determining resistor and of value Rq=RQp. The two capacitors are equal and of value C=1/WpR. each. The two equal banks of capacitors are used to control ωp. Each bank contains n binary weighted capacitors connected in series through analog CMOS switches as shown in Fig.(3.5). Using a digital binary word of n bits to control Wp, 2^n different values of C will result at the 2 terminals of both capacitors banks that correspond to 2^n different values of p. Using a similar technique the value of Rq can be controlled through a bank of m binary weighted resistors in series, though analog CMOS switches as shown in Fig.(3.6). Using a digital binary word of m bits to control Qp, 2^m different values of Rq can be achieved that correspond to 2^m different values of Qp. Thus, full independent control of the pole pair ωp and Qp are achieved by programming the switches to obtain the corresponding C and Rp. It can be easily shown that with minor modifications, an additional programmable element can be added for the control of the notch frequency.

C. THE REALIZED GIC PROGRAMMABLE FILTER

A complete circuit diagram of the constructed GIC filter is shown in Fig.(3.4). The values of m and n were selected to m=n=4. Thus, 15 different values of ωp (fp) and Qp were obtained as it is illustrated at the corresponding Table (3.3) and (3.4). the (designed) banks of the resistors for the control of Qp and the capacitors for the control of ωp (fp) along with their control switches are shown correspondingly in Fig.(3.7) and Fig.(3.8).

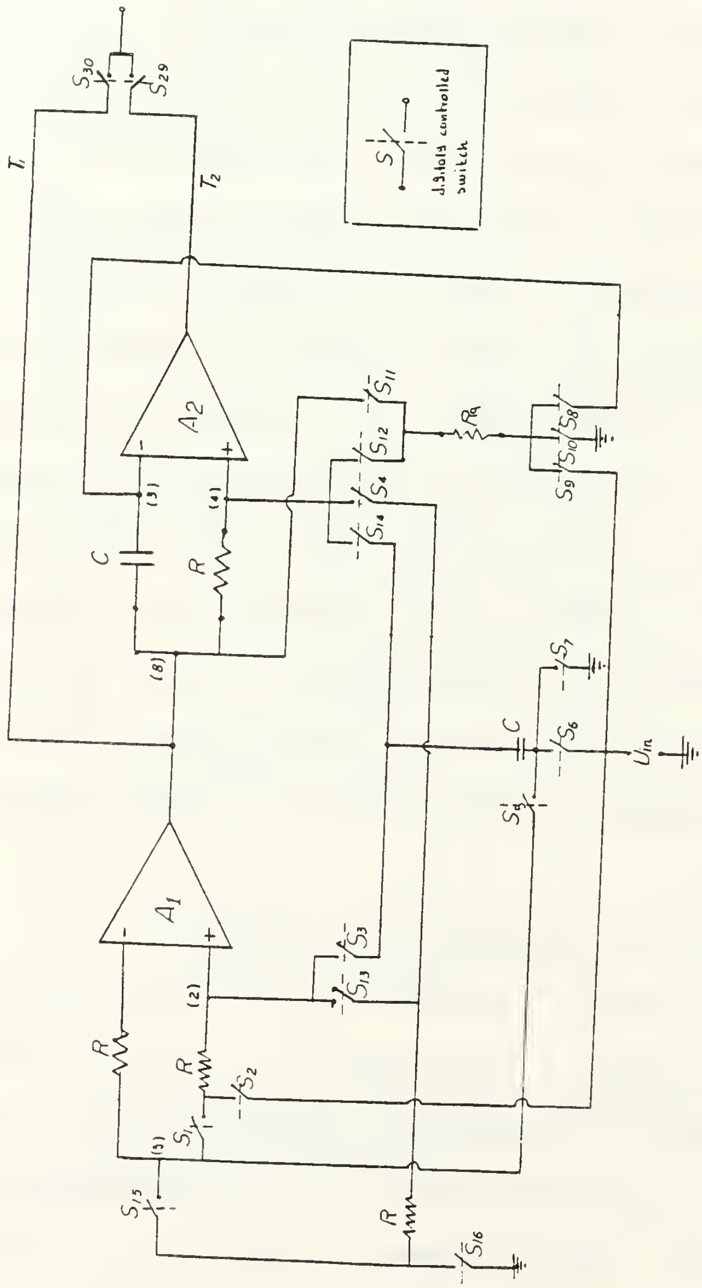


Fig. 3.4 - The Complete Circuit Diagram
of the Programmable GIC Filter

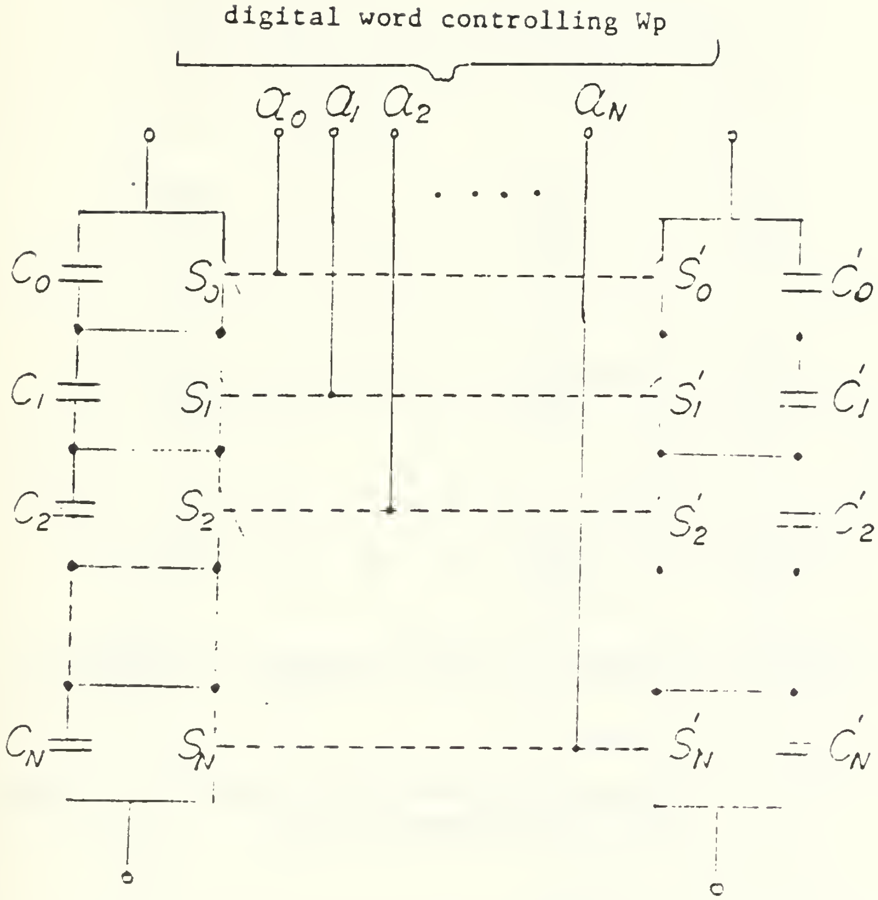
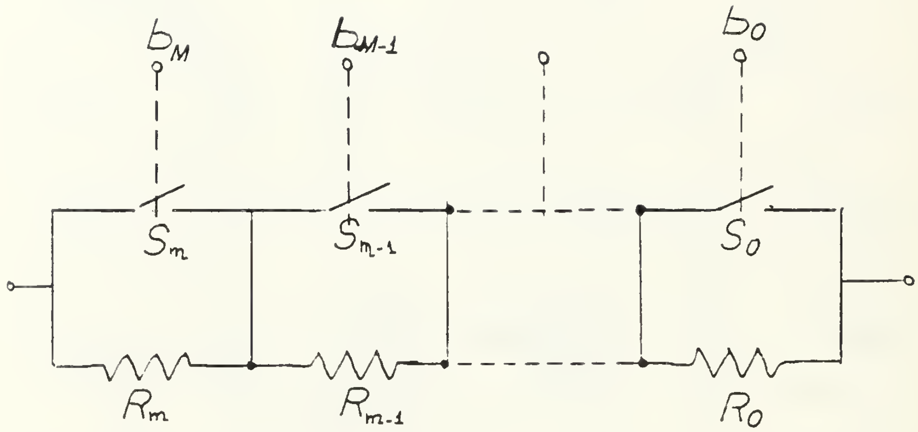


Fig. 3.5 - The Two Capacitor Banks Realizations for the Programming of ω_p .



(For linear Qp control, $R_{j+1}=2R_j$ resulting in $RQp=\sum_{j=0}^{m} R_0 b_j$)

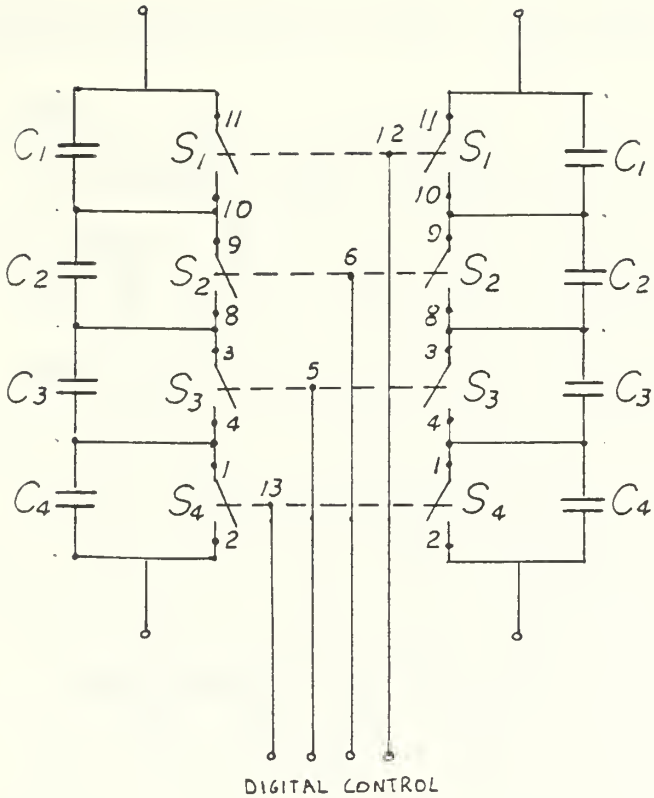
Fig. 3.6 - The Resistor Bank Used to Realize R_q Needed for the Programming of Q_p.

switch control				C	Fp
Sa	Sb	Sc	Sd	nF	Khz
0	0	0	0	5.96	16.693
0	0	0	1	6.34	15.709
0	0	1	0	7.16	13.887
0	0	1	1	7.72	12.875
0	1	0	0	8.42	11.810
0	1	0	1	9.20	10.856
0	1	1	0	10.00	10.043
0	1	1	1	11.00	9.047
1	0	0	0	15.10	6.650
1	0	0	1	17.60	5.637
1	0	1	0	20.80	4.823
1	0	1	1	26.00	3.828
1	1	0	0	35.50	2.805
1	1	0	1	55.00	1.810
1	1	1	0	100.00	.995

Table 3.3 The four - bit words that control Fp and the corresponding capacitor

switch control				Rq	Qp
Sa	Sb	Sc	Sd	K.	
0	0	0	0	24.0	15
0	0	0	1	22.4	14
0	0	1	0	20.8	13
0	0	1	1	19.2	12
0	1	0	0	17.6	11
0	1	0	1	16.0	10
0	1	1	0	14.6	9
0	1	1	1	12.8	8
1	0	0	0	11.2	7
1	0	0	1	9.6	6
1	0	1	0	8.0	5
1	0	1	1	6.4	4
1	1	0	0	4.8	3
1	1	1	1	3.2	2
1	1	1	0	1.6	1

Table 3.4 The four - bit words that control Rp and Qp



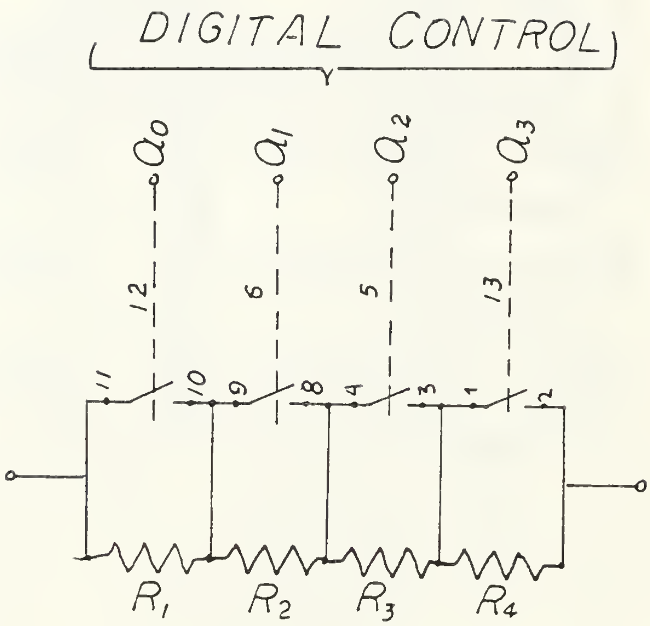
C_1 = 100uf

C_2 = 55uf

C_3 = 25uF

C_4 = 11uF

Fig. 3.7 - The Constructed Blocker Capacitor Controlling the fp.



R_1 = 1.6k

R_2 = 3.2k

R_3 = 6.4K

R_4 = 12.8k

Fig. 3.8 - The Constructed Block of Resistance Controlling Qp.

IV. COMPUTER SIMULATION OF THE PROGRAMMABLE FILTER

A. INTRODUCTION

In order to observe the theoretical responses of the different realizations as well as to compare them with the experimental measurements, two computer programs were written in Fortran. those programs are shown in Appendices A and B. The first one simulates the programmable filter's responses T1(s), T2(s) and T3(s) as functions of the network's admittances Yi, i=1 . . .8, and the "constant" Op.Amps. gains A1 and A2. A realistic value of $A2=A2=10^5$ was given to the above gains corresponding to 741 Op. Amps. used in the experiment. the second program simulates the programmable filter's responses T1(s), T2(s) and T3(s) as functions of the network's admittances Yi, i=1 . . . 8 and the frequency dependent Op. Amps. gains A1 and A2. In this second case a single pole approximation value of wils was assigned for Ai(i=1, 2). Where, Wi is the GBWP of the Op. Amps. used, a value of $W1=W2=2\pi\,x10^6$ was given to the above Wi(i=1, 2) corresponding to 741 Op. Amps. used throughout the research.

To be able to compare the computer simulation results with the experimental ones obtained, the same values of admittances were used as input data. That means that the same values of R, Rq and C according to Chapters III and V were used. With the value of R selected to be 1.6k Table

(4.1) illustrates the different values of Rqs used in the simulation to realize the different values of Qp and Table (4.2) illustrates the different values of capacitors used in the simulation to realize the different frequencies Wps.

The "DO CASE I" command of Fortran simulated the digital logic (including the control switches) used to realize the different responses of the programmable filter that is LP, HP, BP, N and AP. The frequency's (ωp) translations and the different QP values were simulated by changing the values of Rq and Cs at every run of the program.

The above programs also simulated the transfer function of two cascaded GIC programmable filters as it will be discussed later in Chapter VI. Each of the filters could have been at different realization (as well as at different p and Qp) relative to the other one. The 25 possible combinations of transfer function realizations are shown in Table 4.3.

B. SIMULATION RESPONSE(S)

1. Low Pass (LP) Realization

Using the elements values prescribed in Table (3.1) yields the following L.P. transfer function:

$$T_2(s) = \frac{2\omega_p^2}{s^2 + \frac{\omega_p}{Q_p} s + \omega_p^2} \qquad (4.1)$$

For R=1.6kΩ	
Rq kΩ	Qp
1.1	0.7
1.6	1.0
3.2	2.0
6.4	4.0
11.2	7.0
16.0	10.0
32.0	20.0

Where Qp = $\underline{Rq}$

Table 4.1 - The Resulting Qp Values for the Different Introduced Rq Values in the Computer Simulation.

For R-1.6kΩ

C nf	fp khz
100.00	.099
50.00	1.99
26.00	3.83
12.5	7.96
6.6	15.10
3.3	30.16
0.85	117.09
0.42	236.67

$$fp = \frac{1}{2\pi \cdot R.C.}$$

Table 4.2 - The Resulting ωp Values for the Different Introduced C Value (R constant 1.6kΩ)

FILTER 1	FILTER 2	J	I
LP	LP	2	2
HP	HP	1	1
BP	BP	4	4
N	N	3	3
AP	AP	3	3
LP	HP	2	1
LP	BP	2	4
LP	N	2	3
LP	AP	2	3
HP	LP	1	2
HP	BP	1	4
HP	N	1	3
HP	AP	1	3
BP	LP	4	2
BP	HP	4	1
BP	N	4	3
BP	AP	4	3
N	LP	3	2
N	BP	3	1
N	BP	3	4
N	AP	3	3
AP	LP	3	2
AP	HP	3	1
AP	BP	3	4
AP	N	3	3

Table 4.3 - The 25 Possible Combinations of Transfer Function Realizations

This transfer function assumes the following complex value

$$T_2(j\omega_p) = -2jQ_p \tag{4.2}$$

with magnitude of

$$|T_2(j\omega_p)| = 2Q_p = 20\log|2Q_p| \ , db \tag{4.3}$$

Fig. 4.1 illustrates the theoretical "ideal" LPF magnitude response for fp=3.8Khz and for 3 different values of Qp. The simulation results match the equations of (4.2).

Fig. 4.2 illustrates the theoretical "ideal" LPF magnitude response for Qp=2 and for 3 different values of ωp. the Op. Amps. gains Ai, (i=1, 2) are frequency depended. This dependance affects the magnitude of the filter and causes a frequency shift from the theoretical value of the ideal's case. Figs.(4.3) and (4.4) illustrates the ideal vs. nonideal theoretical LPF amplitude responses.

Data extracted from Figs.(4.3) and (4.4) are illustrated in Tables (4.4(a)) and (4.4(b)) simultaneously.

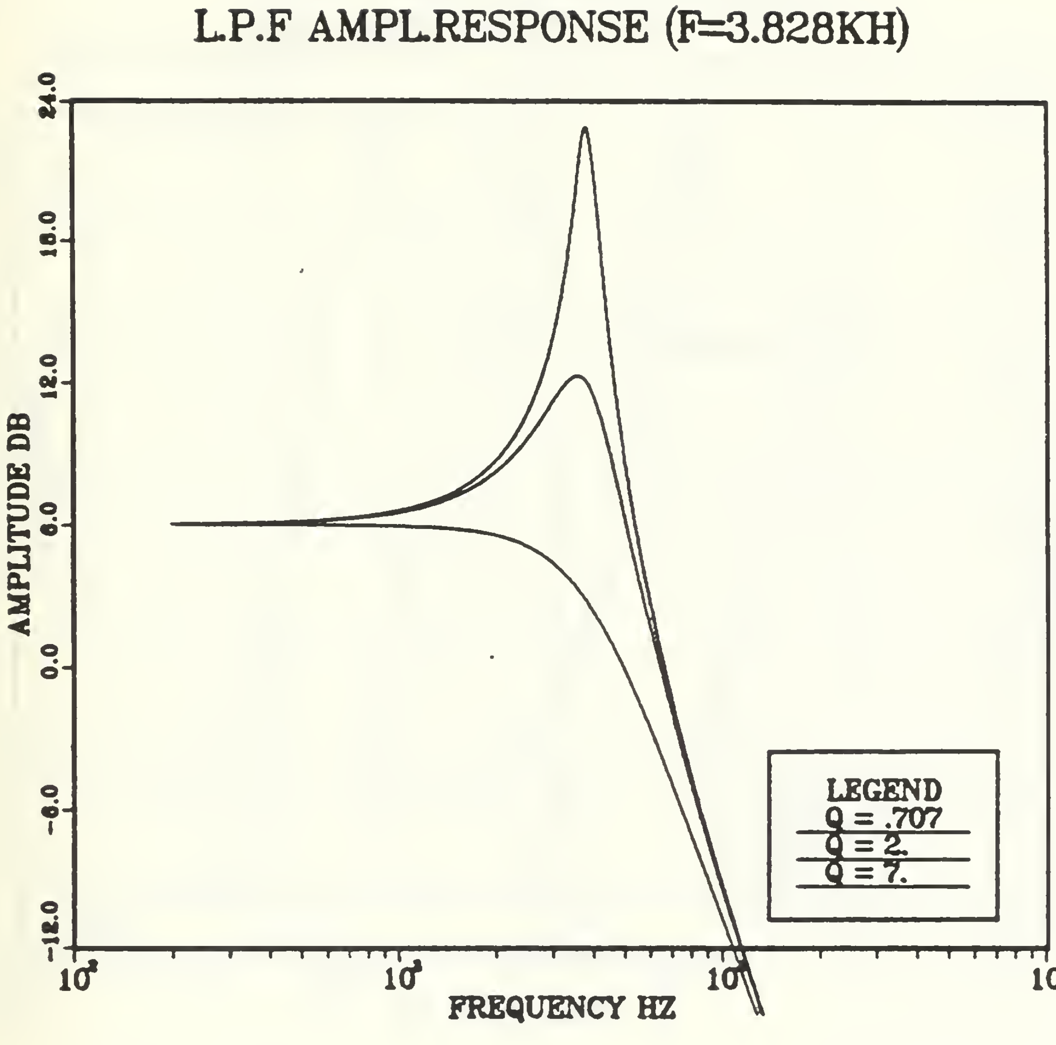


Fig. 4.1 - "Ideal" LPF Amplitude Response

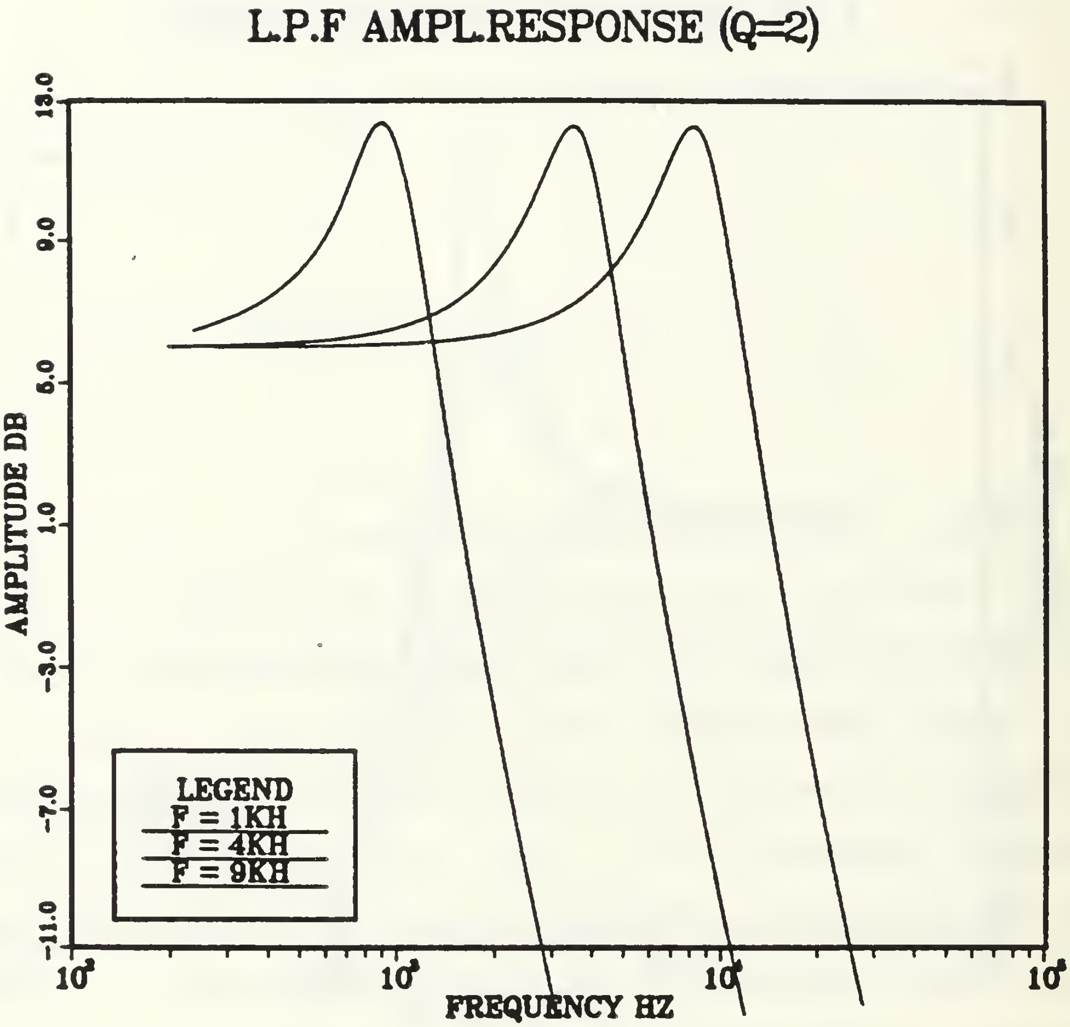
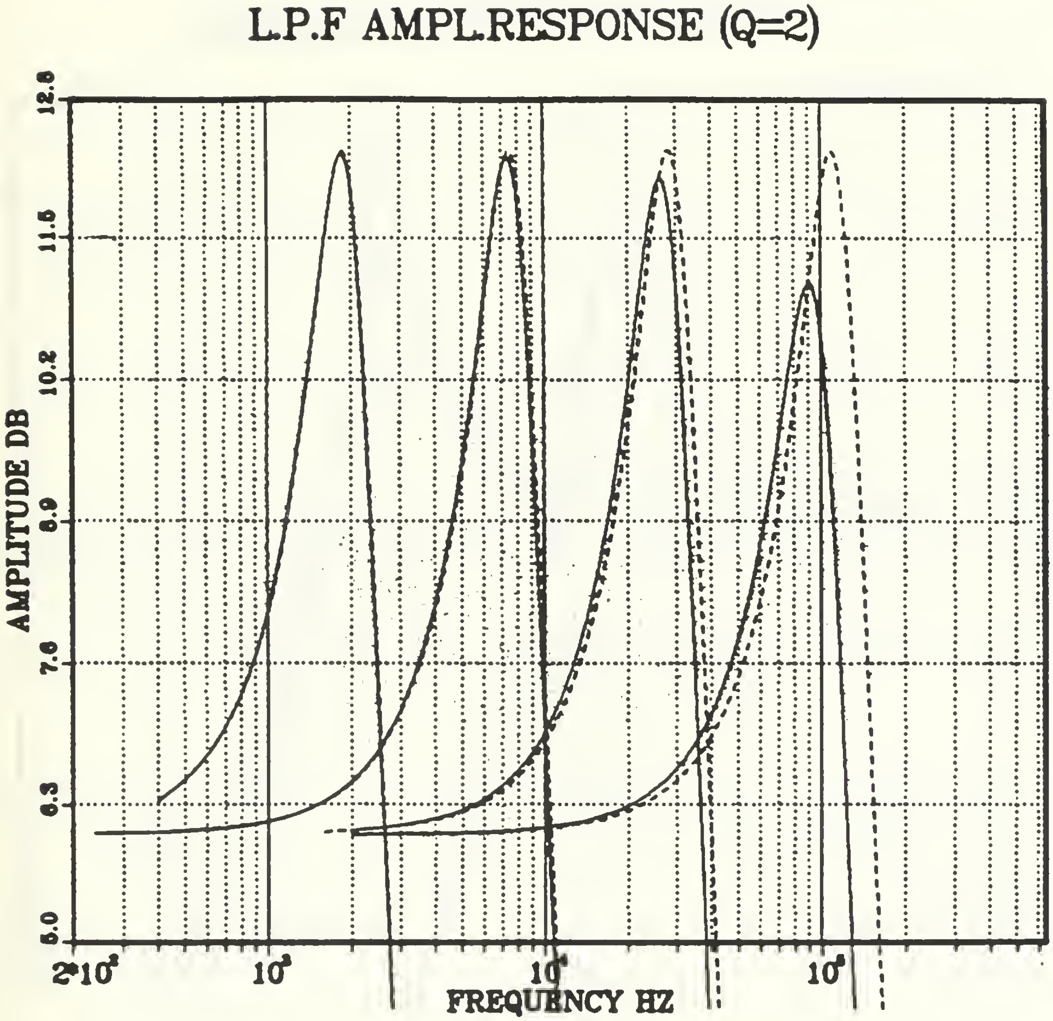


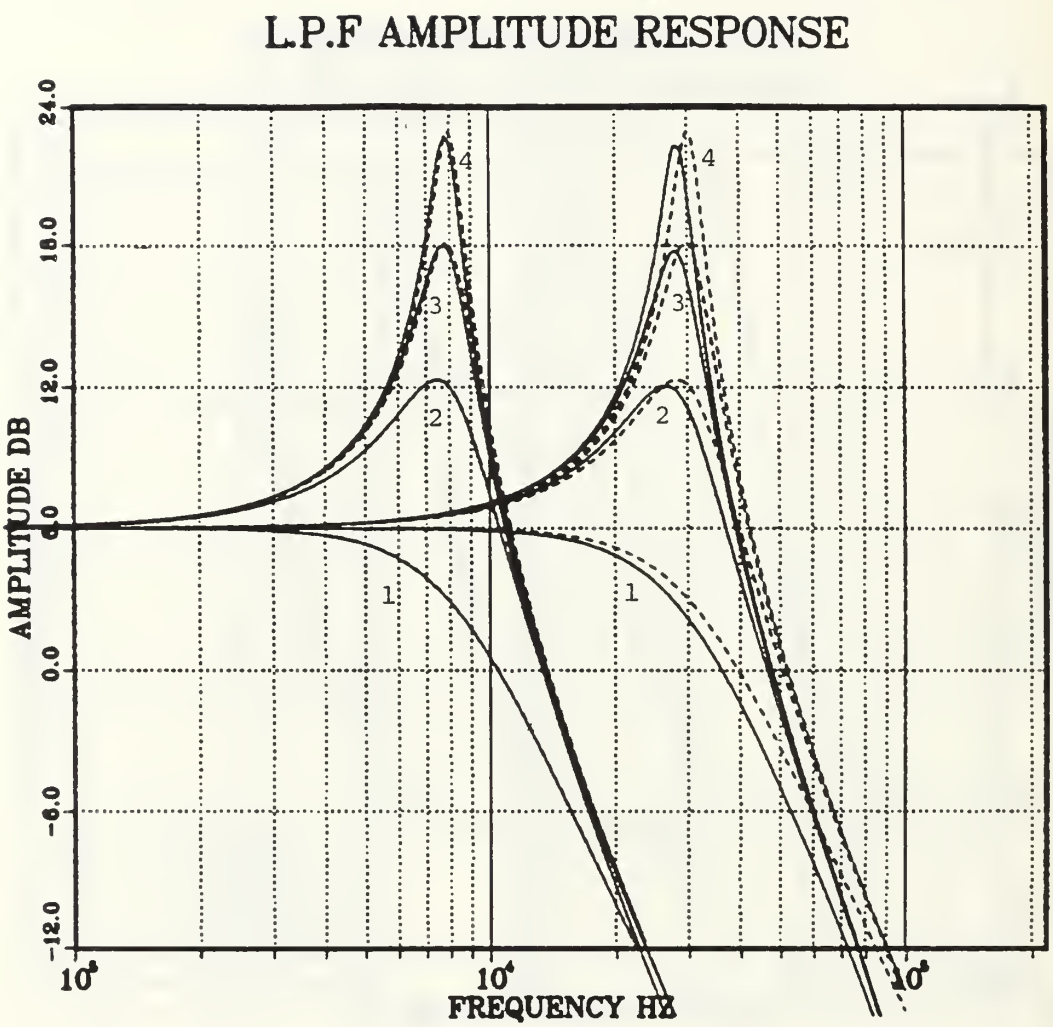
Fig. 4.2 - "Ideal" LPF Amplitude Response



------ ideal

______ nonideal

Fig. 4.3 Ideal .vs. nonideal L.P.F. amplitude response for Q=2 and frequencies (1.99K, 7.96K, 30.16K and 117.1K)hz



1. Q = 0.7
2. Q = 2.0
3. Q = 4.0
4. Q = 7.0

Fig. 4.4 Ideal .vs. nonideal L.P.F. amplitude responce for frequencies (7.96K, 30.1K)hz and variety of Qs.

Data From Fig. (4.4) LPF Characteristic

Ideal			Nonideal	
Qp	$\lvert T_2(s)\rvert$ at pick value	Rq kΩ	Qp	$\lvert T_2(s)\rvert$ at pick value
.07	3.0	1.1	6.7	3.0
2.0	12.0	3.2	2.0	12.0
4.0	18.1	6.4	4.0	18.1
7.0	22.9	17.2	7.0	22.7
The pick values obtained at fpick = 7.9k			The pick values obtained at fpick - 7.9k	

(b)

Ideal		C	Nonideal	
Fp khz	Amp. Response dB	nf	fp khz	Amp. Response dB
1990.45	12.04	50	1990.45	12.72
7165.60	12.24	12.5	8359.87	12.113
28662.42	12.31	6.6	25875.80	12.04
109745.50	12.31	0.085	90565.00	11.08

Table 4.4 - Data from Figs. (4.3) and (4.4) indicating the Affect of Frequency Dependency of Ai, i=1, 2

2. <u>High Pass (HP) Realization</u>

Using the elements value shown in Table (3.1) yields the following HP transfer function:

$$T_1(s) = \frac{2s^2}{s^2 + \frac{\omega_p}{Q_p} \cdot s + \omega_p^2} \tag{4.4}$$

takes the following complex value at ω_p.

$$T_1(j\omega_p) = 2jQ_p \tag{4.5}$$

which magnitude is

$$|T_1(j\omega_p)| = 2Q_p = 20\log(2Q_p)\text{ , db} \tag{4.6}$$

Fig. (4.5) illustrates the theoretical "ideal" HPF amplitude response for fp=3.8KHZ and for 3 different values of Qp. The computer simulation results match that of the (4.6) relation.

Fig. (4.6) illustrates the theoretical "ideal" HPF amplitude response for Q=2 and for 3 different frequencies.

Fig. (4.7) illustrates how this dependency effects the amplitude of the HP filter. (The amplitude decreases as

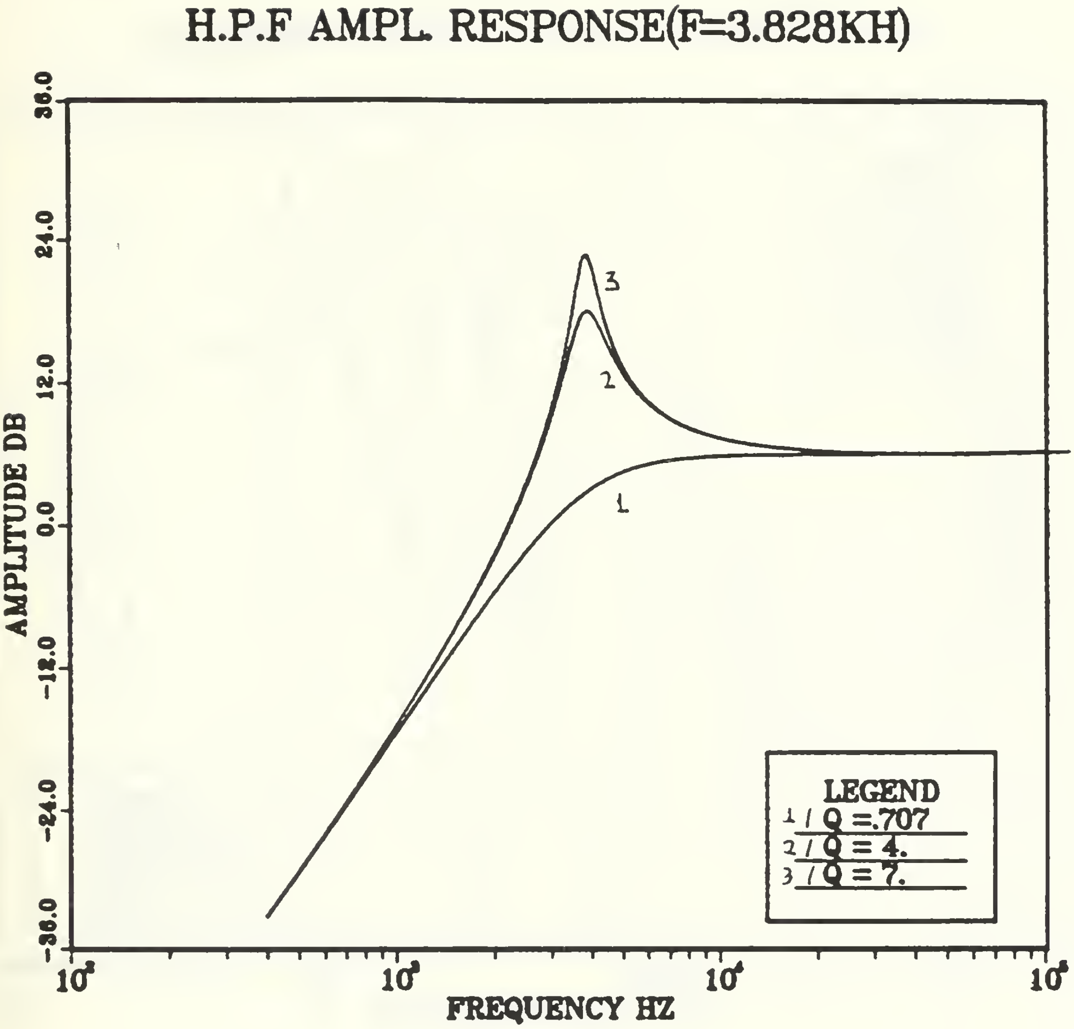


Fig. 4.5 - "Ideal" HPF Amplitude Response for a Variation of Qs.

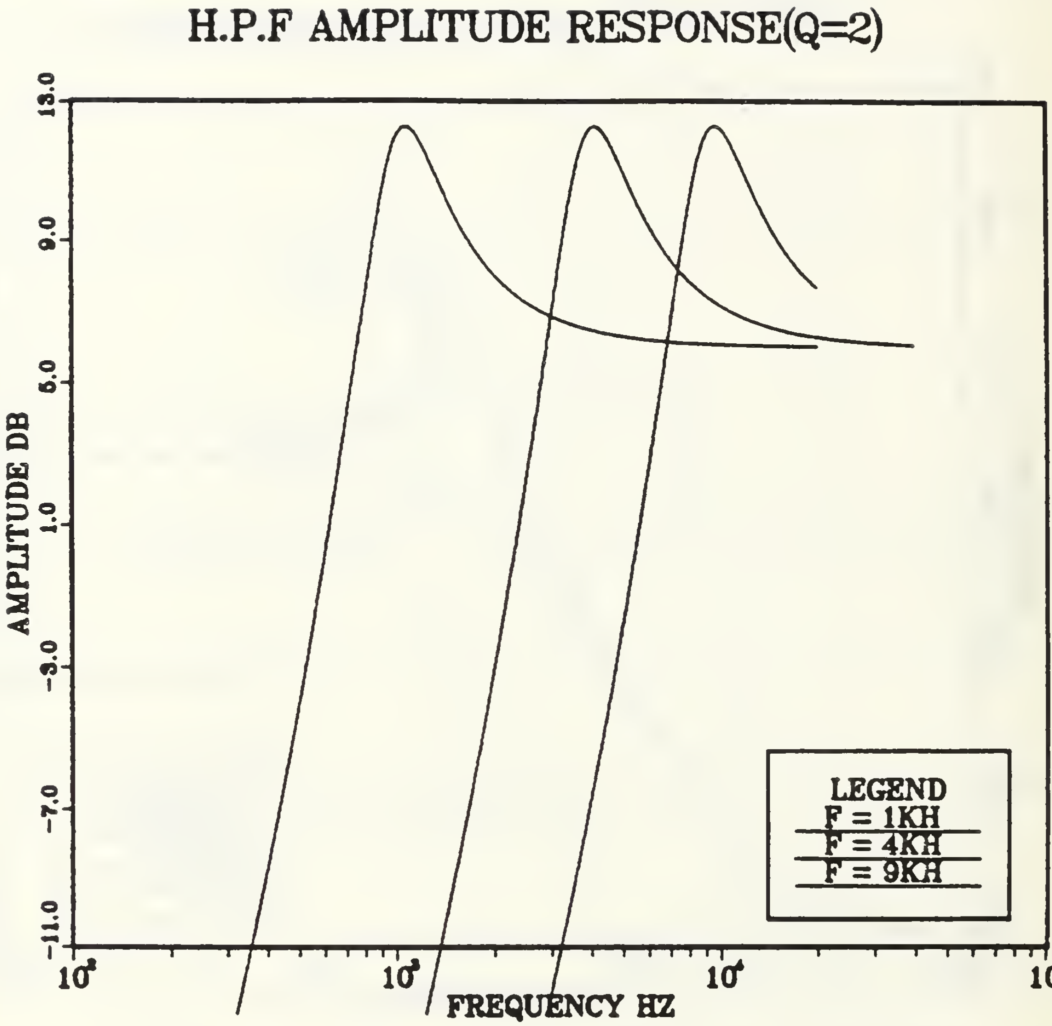


Fig. 4.6 - "Ideal" HPF Amplitude Response.

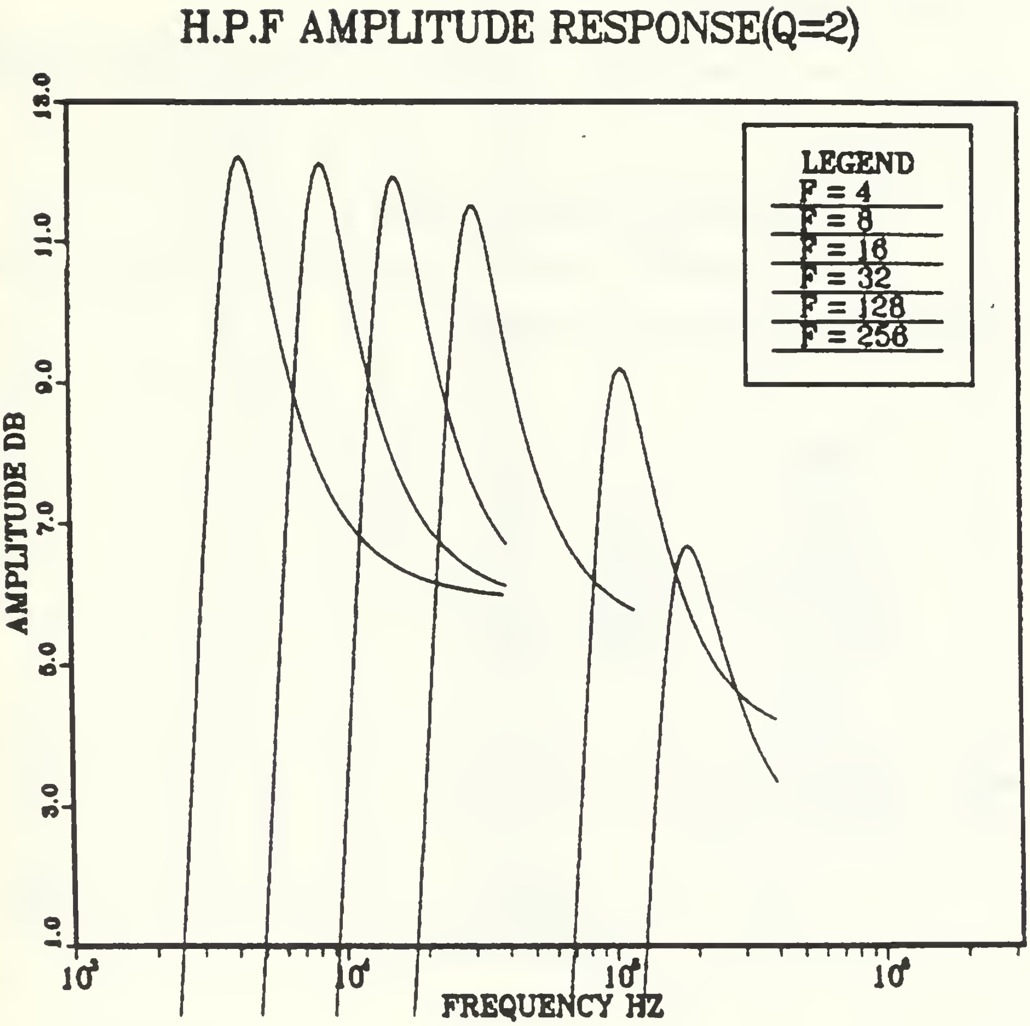


Fig. 4.7 - Nonideal HPF Amplitude Response.

the frequency increases) while Figs. (4.8) and (4.9) illustrate the frequency shift of the amplitude response due to that dependency.

From the date of Fig. (4.8), Table (4.5) were constructed.

3. <u>Band Pass (BP) Realization</u>

Using element values from Table 3.1 the following BP transfer function is achieved:

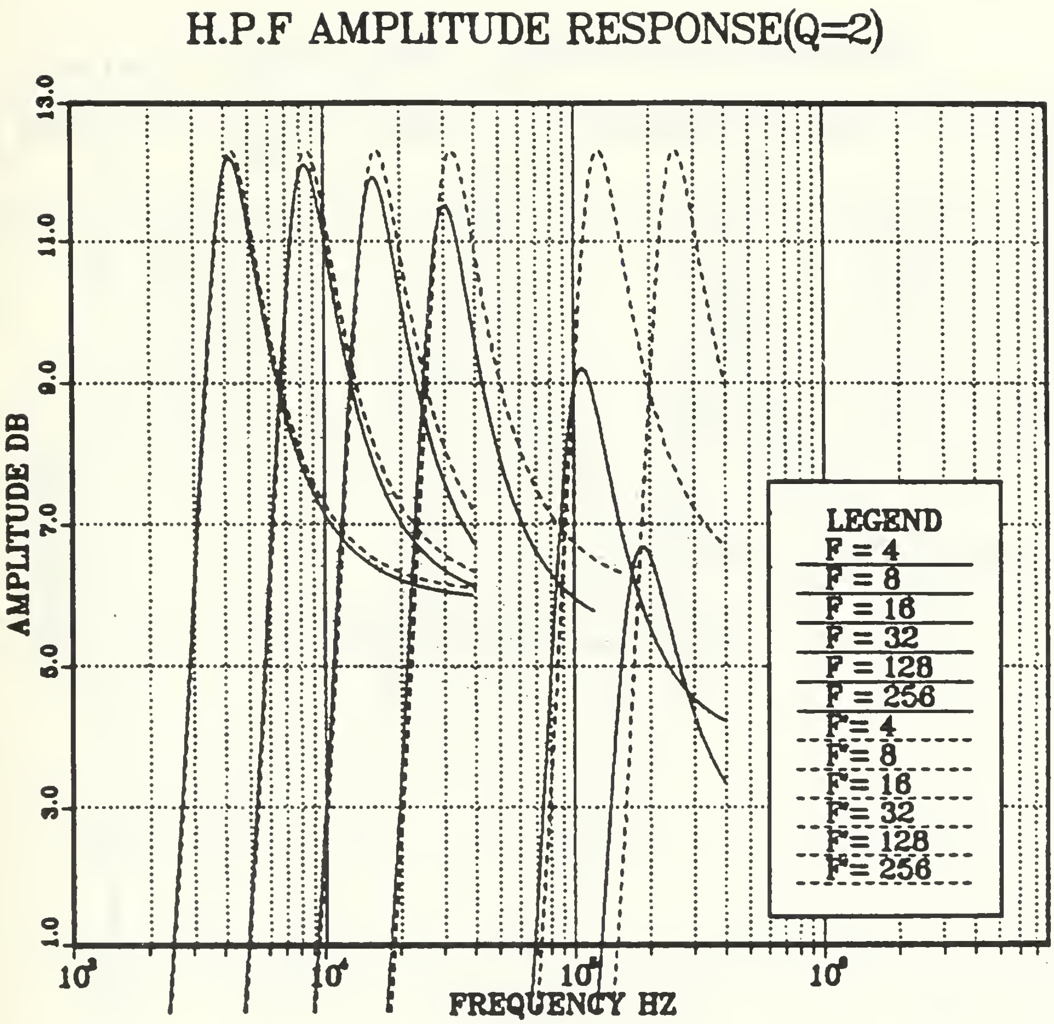
$$T_1(s) = \frac{2\left(\frac{\omega_p}{Q_p}\right)s}{s^2 + \frac{\omega_p}{Q_p}s + \omega_p^2} \qquad (4.7)$$

takes the following value at ωp.

$$T_1(j\omega_p) = 2 \qquad (4.8)$$

which has constant magnitude of 6dB

$$|T_1(j\omega_p)| = 2 = 20\log 2 = 6db \qquad (4.9)$$

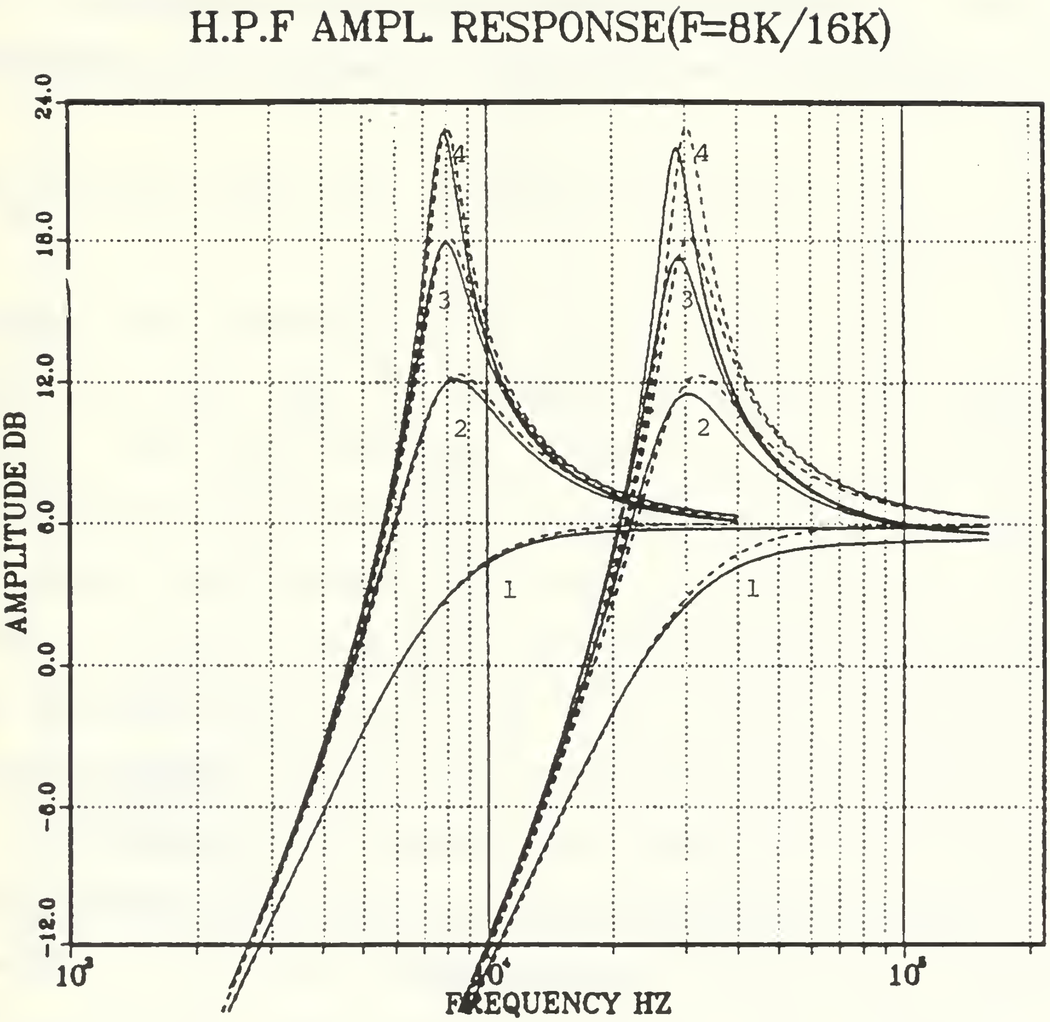


----- Ideal
______ nonideal

Fig. 4.8 - "Ideal" vs.Nonideal HPF Amplitude Response.

Ideal		c	Nonideal	
fp Hz	Amp Response dB	nf	fp Hz	Amp Response Db
4,378.98	12.28	26	4,378.98	12.148
8,359.87	12.30	12.5	8,359.87	12.11
15,923.57	12.31	6.6	15,923.57	12.02
31,847.13	12.31	3.3	31,847.73	11.41
123,407.00	12.31	0.85	107,484.00	9.21
252,781.05	12.31	0.43	187,101.80	9.21
242,781.05	12.32	0.43	187,101.80	6.69

Table 4.5 - Data Illustrating the Ideal vs. the Nonideal Responses of the HPF



1. Q = 0.7
2. Q = 2.0
3. Q = 4.0
4. Q = 7.0

Fig. 4.9 - "Ideal" vs. Nonideal HPF Amplitude Response.

Fig. (4.10) illustrates the theoretical BPF magnitude response for fp=3.83 KHZ for different values of Qp. These agree with the (4.9) equation since as it is indicated by the simulation plot the amplitude is constant and independent of Qp.

Fig. (4.11) illustrates the above concept but at fp=15.1KHZ.

Fig. (4.12) and (4.13) illustrate the theoretical "ideal" BPF amplitude response for Qp=2 and for different values of p. As it is indicated from Fig. (4.13) the amplitude remain constant even at very high frequencies (10 HZ). But with Ai, (i=1, 2) depending on frequency the amplitude decreases as the frequency increases. This is indicated in Figs. (4.14) and (4.15) which describe the BPF amplitude response plots for Q=4 and Q=1 respectively and for different frequencies. The frequency dependence of Ai, (i=1, 2) creates a frequency shift from the ideal theoretical value which is indicated in Figs. (4.16) and (4.17). Table (4.6) illustrates the data extracted from figs. (4.16) and (4.17).

4. Notch (N) Realization

Using the admittances value of Table (3.1), the following Notch transfer function can be achieved:

(4.10)

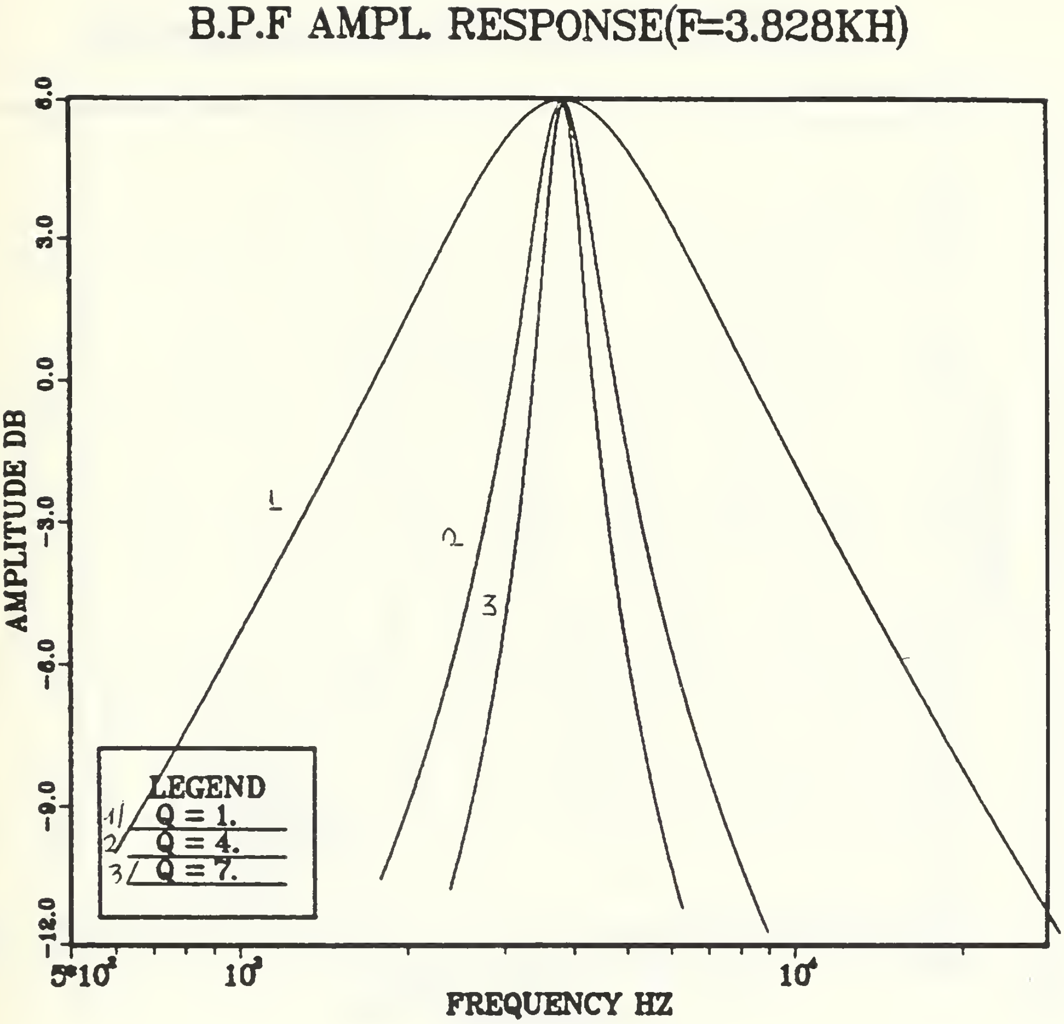


Fig. 4.10 - "Ideal" BPF Amplitude Response.

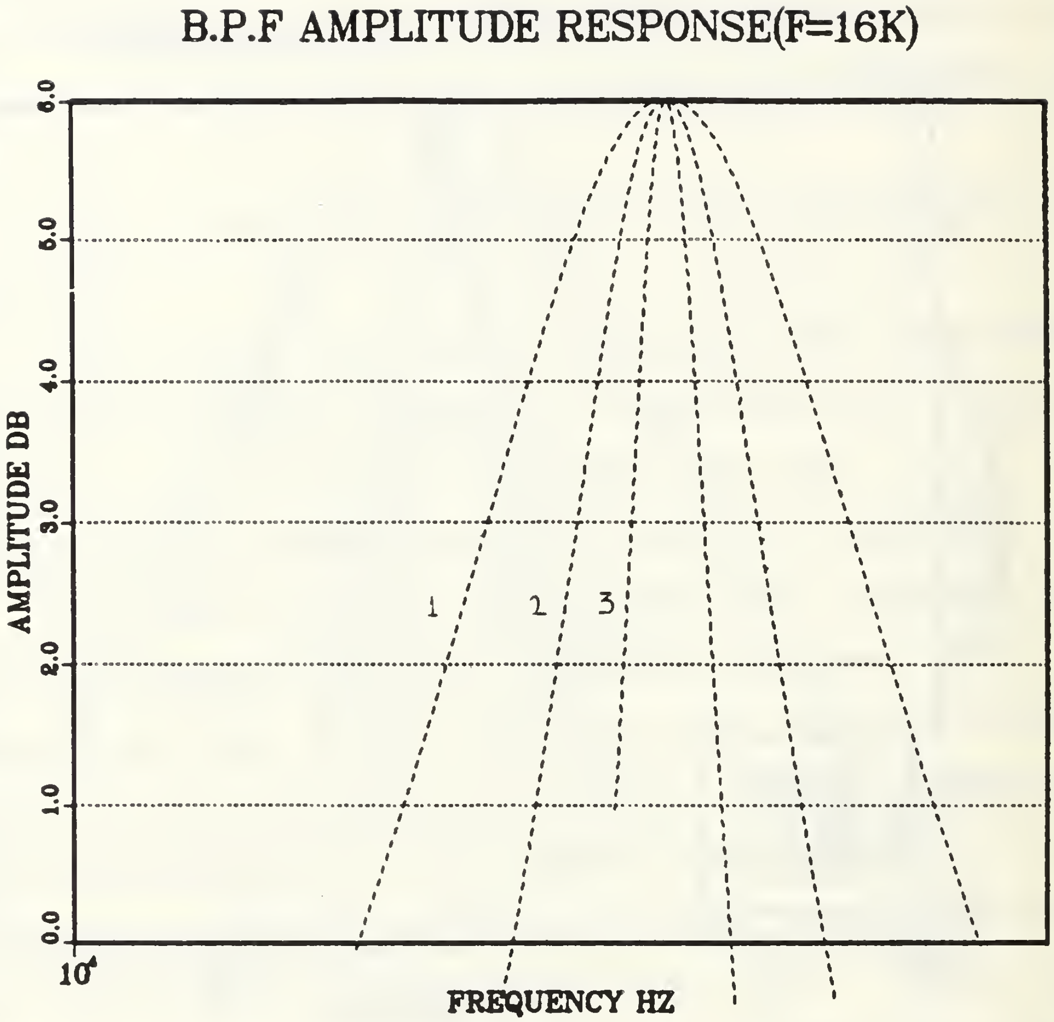


Fig. 4.11 - "Ideal" BPF Amplitude Response.

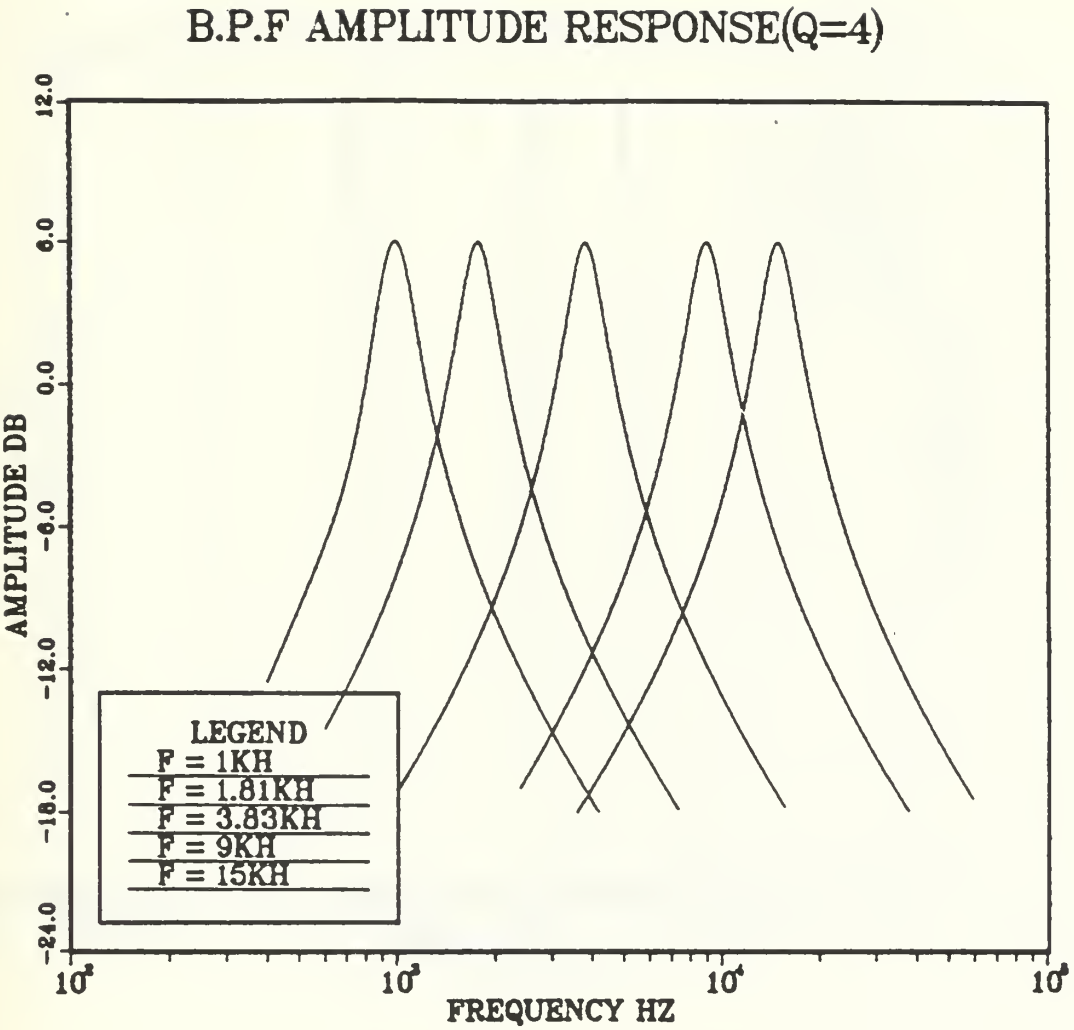


Fig. 4.12 - "Ideal" BPF Amplitude Response.

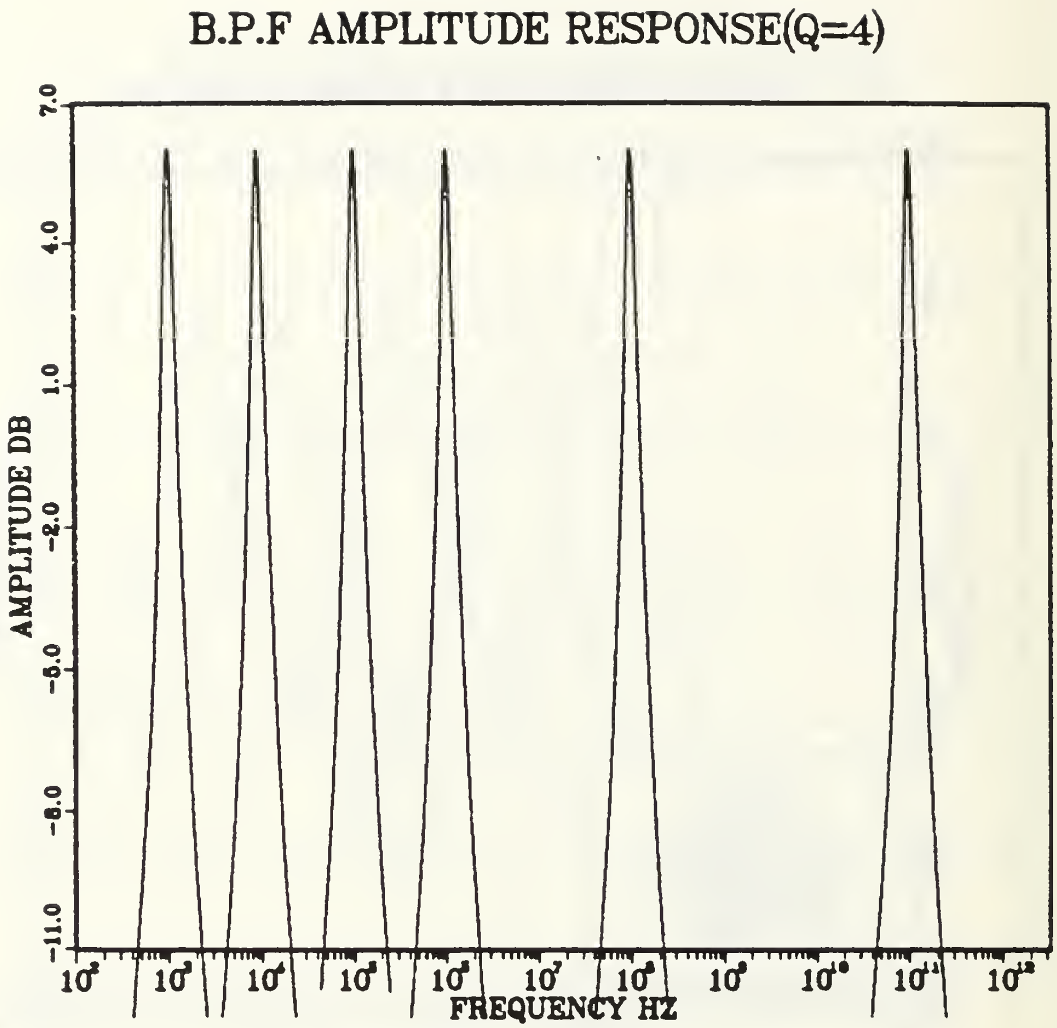


Fig. 4.13 - "Ideal" BPF Amplitude Response.

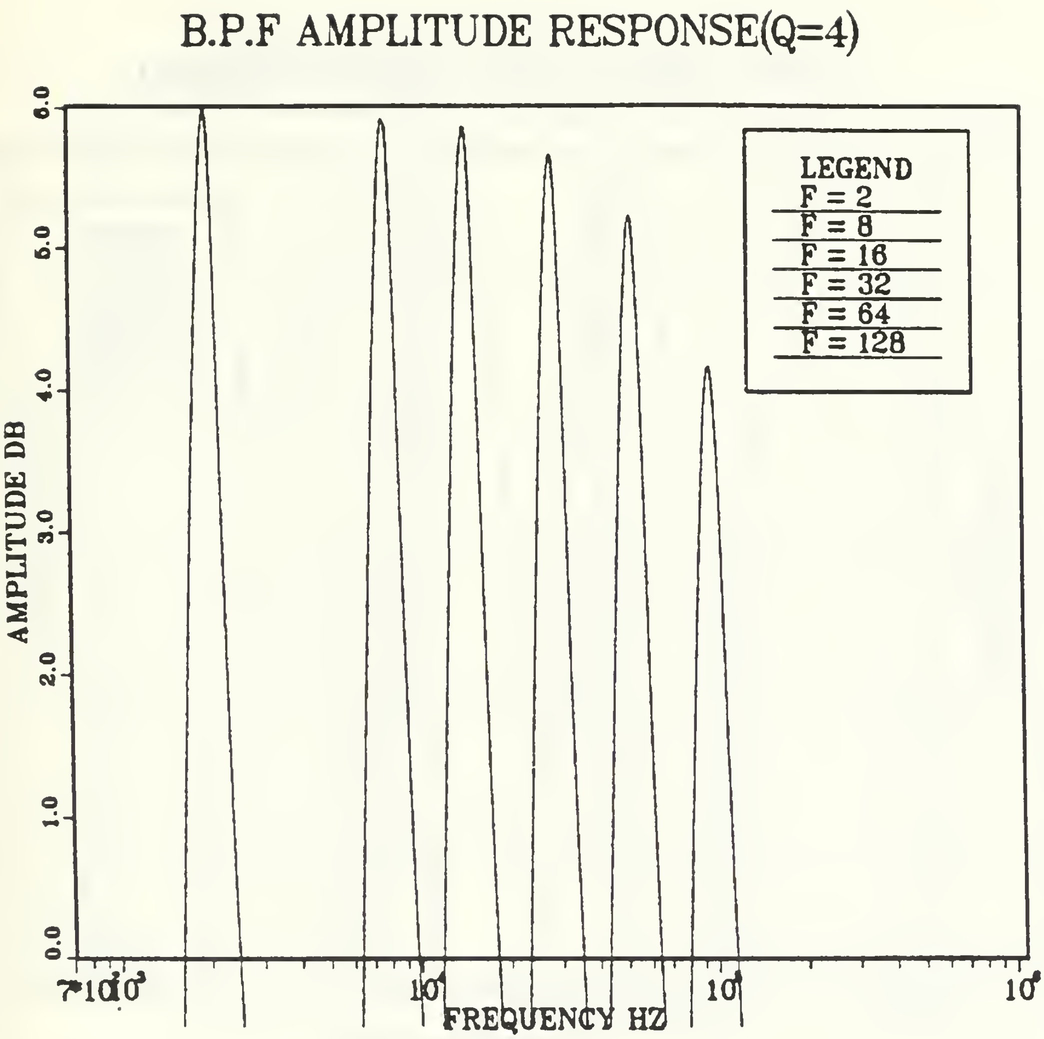


Fig. 4.14 - Nonideal PBF Amplitude Response.

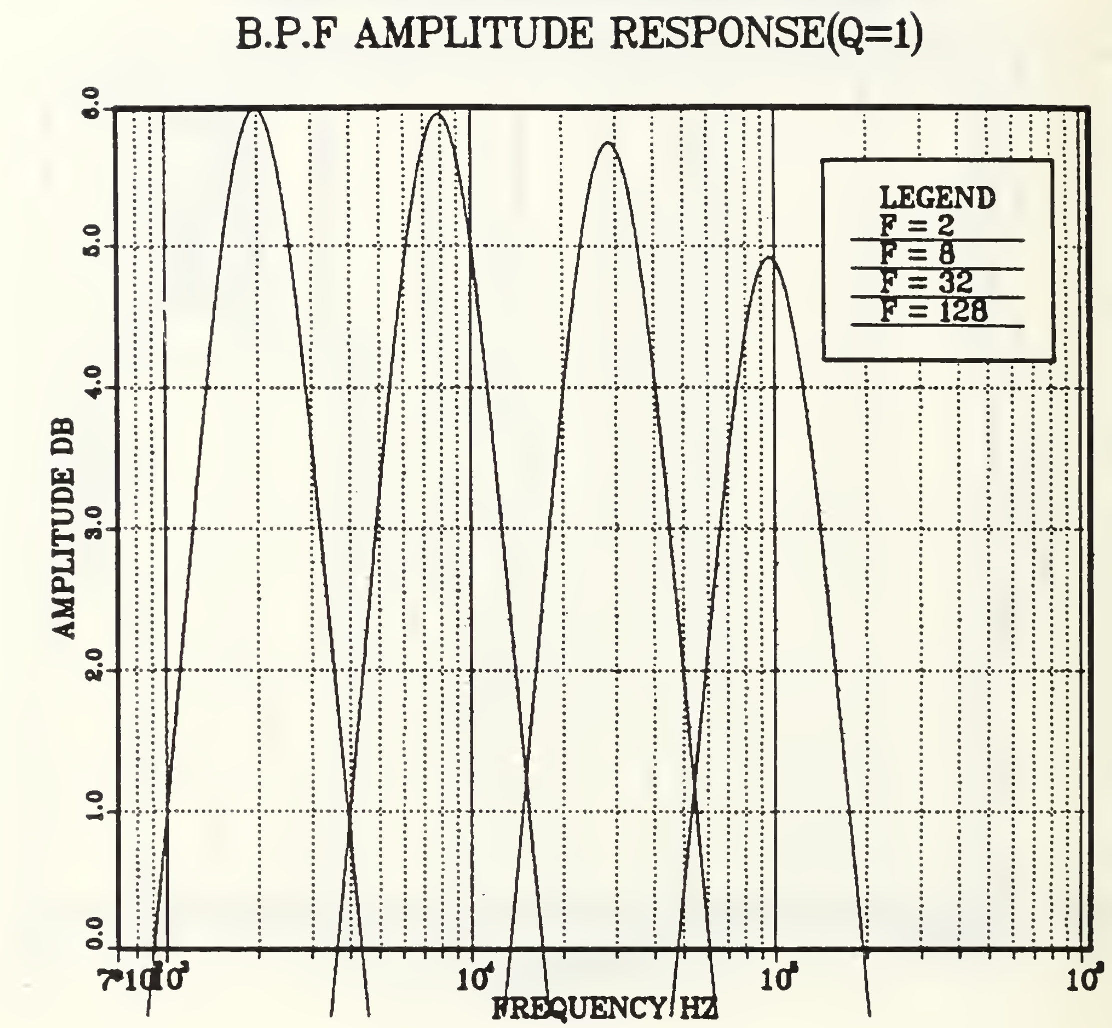


Fig. 4.15 - Nonideal BPF Amplitude Response.

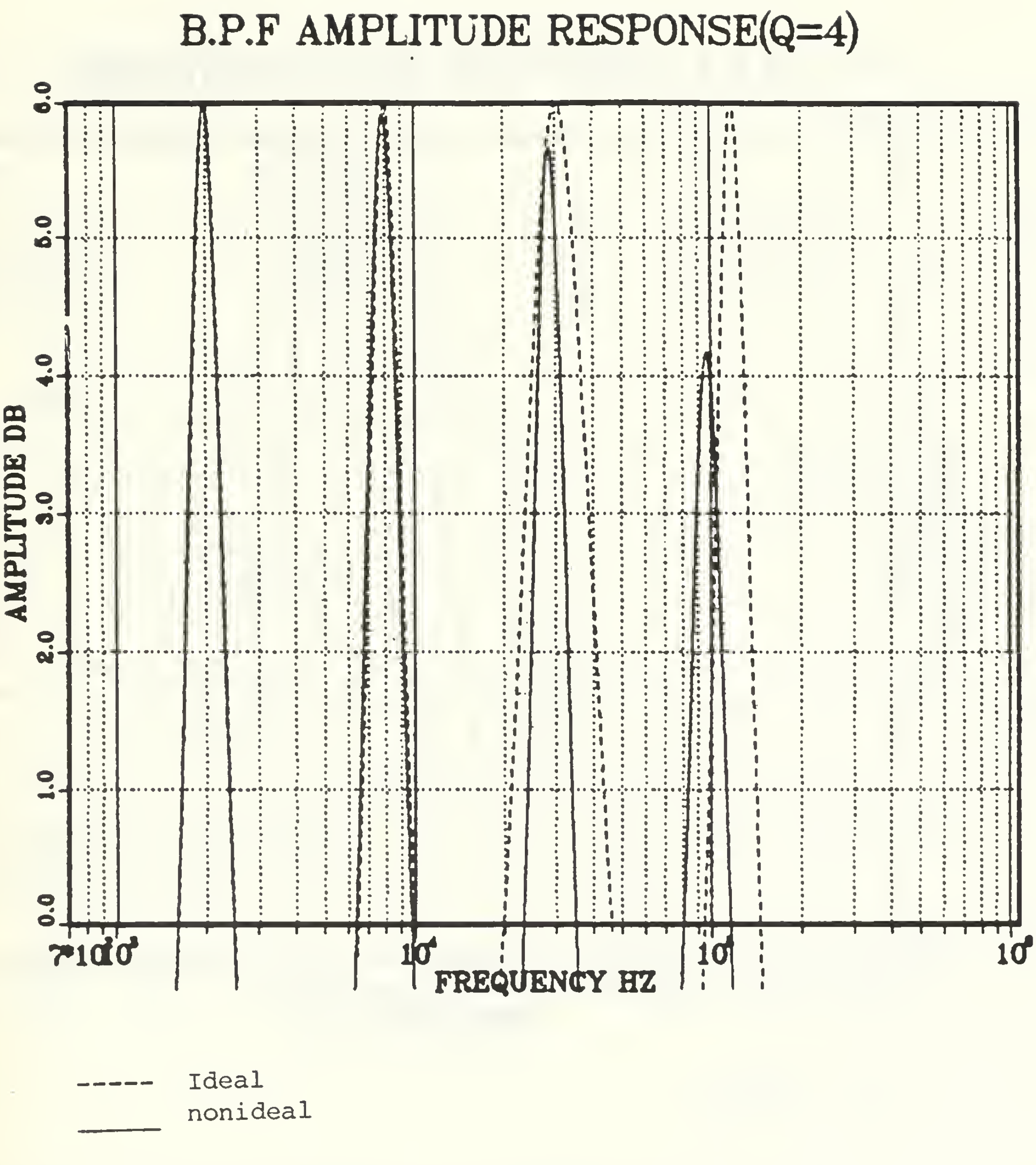
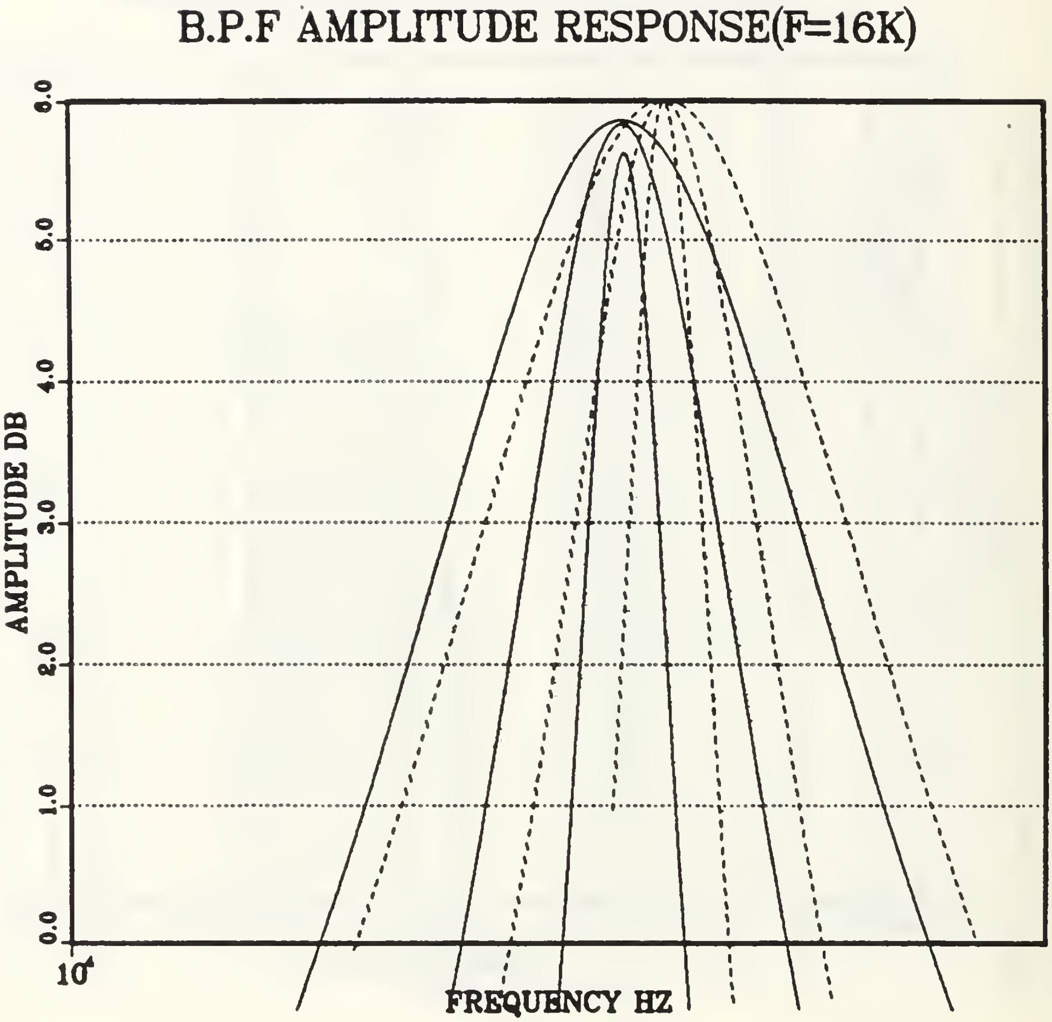


Fig. 4.16 - Ideal vs. Nonideal BPF Amplitude Response.



----- Ideal
______ nonideal

Fig. 4.17 - Ideal vs. Nonideal BPF Amplitude Response.

Ideal		C	Nonideal	
fp khz	Amp. Response dB	nf	fp khz	Amp. Response dB
7,961.78	6.02	12.5	7,969.738	5.958
30,254.78	6.02	3.3	28,662.42	5.75
115,445.80	5.97	0.85	95,541.38	4.92

and

Ideal		Rq	Nonideal	
Qp	Max. Amp. Response dB	k	Qp	Max. Amp. Response dB
4	6.01	6.4	4	5.902
8	6.01	12.8	8	5.728
20	5.95	32.0	20	5.528
fp_1 = 15,127.33Hz			fp_1 = 14,739.29Hz	

Table 4.6 - Data Indicating the Ideal vs. the Nonideal Responses of the BPF Realization

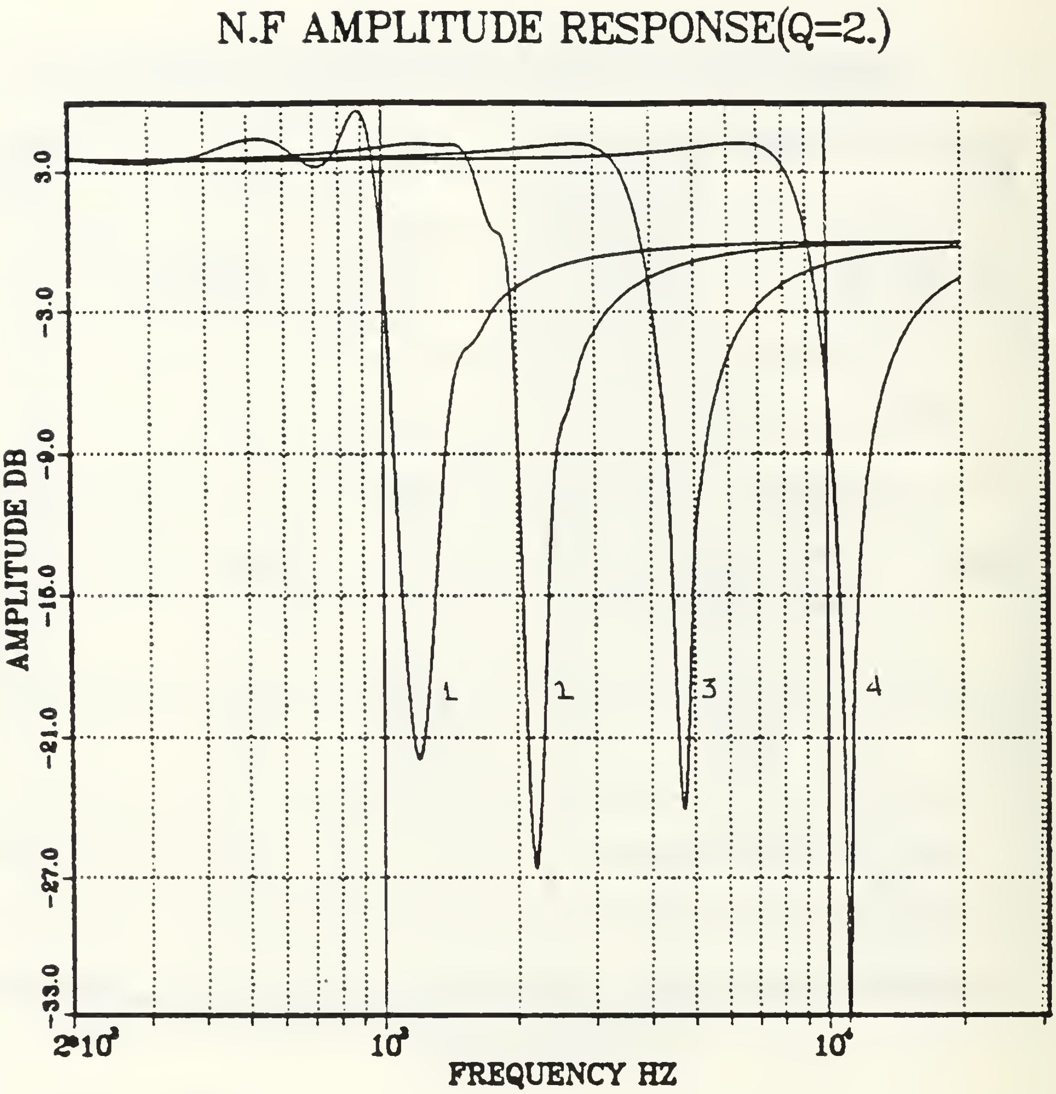


Fig. 4.18 - Ideal Notch Amplitude Response.

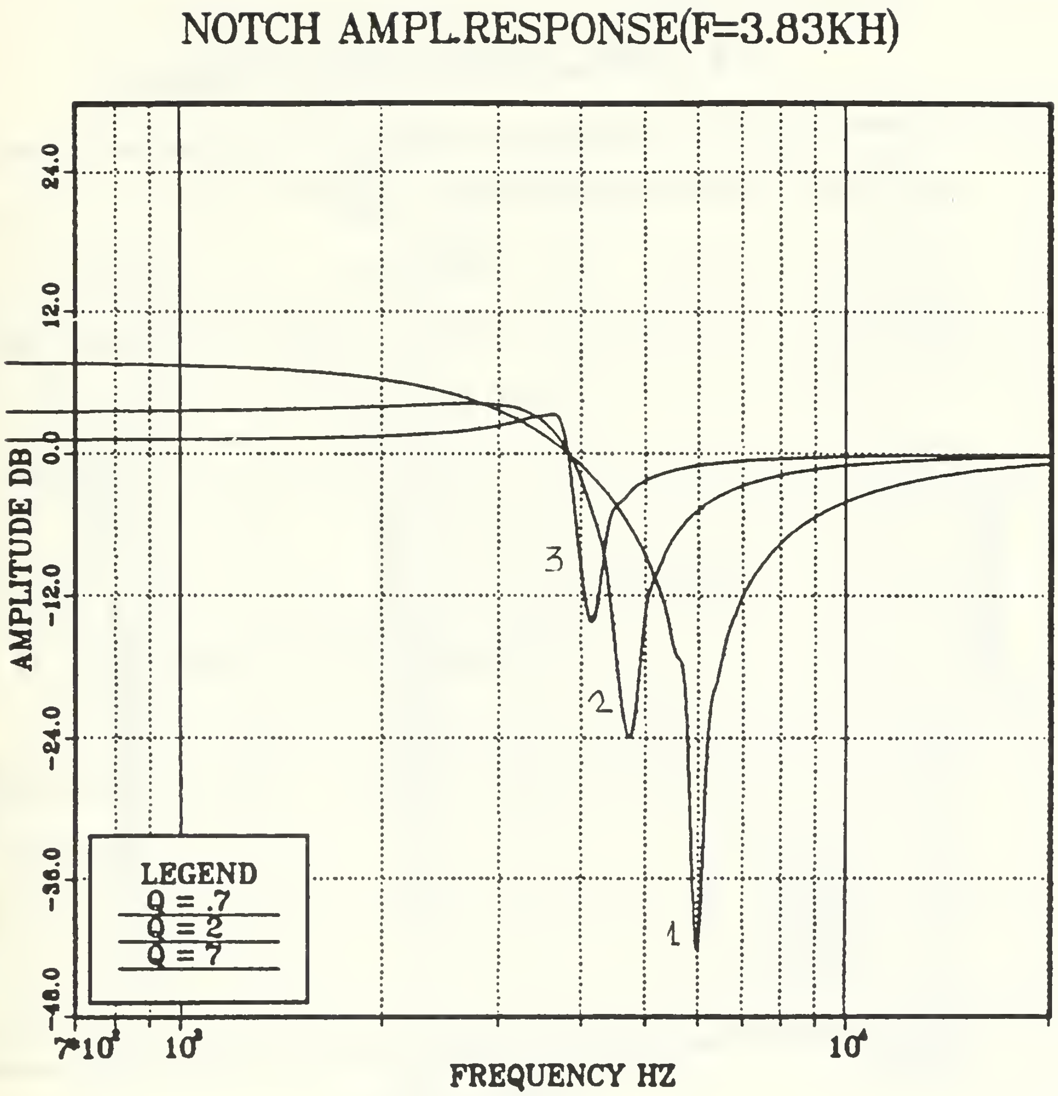
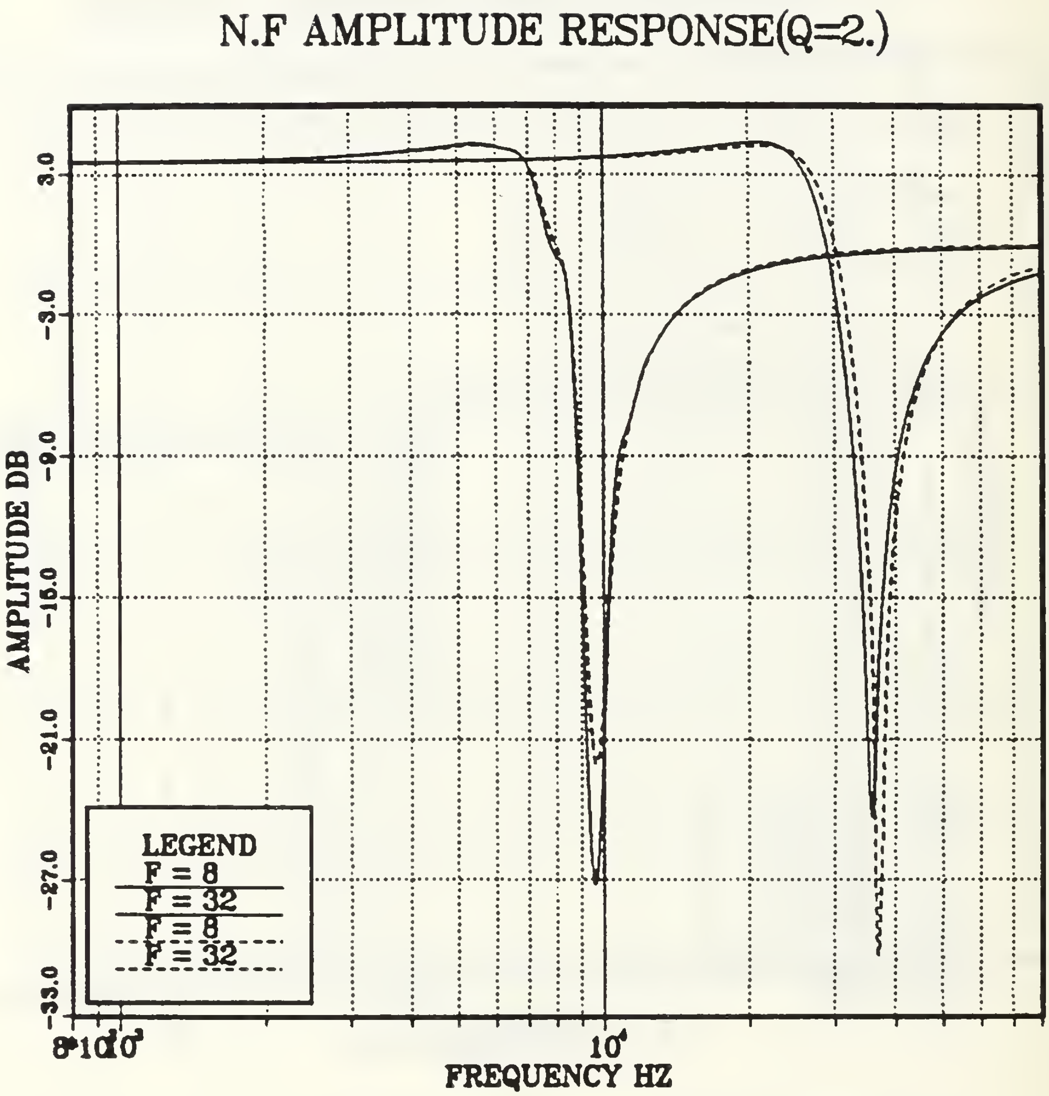


Fig. 4.19 - Ideal Notch Amplitude Response.



------ Ideal
______ nonideal

Fig. 4.20 - Ideal vs. Nonideal Amplitude Response.

where n is the Notch frequency.

At P the transfer function takes the value

$$T_2(j\omega_p) = \frac{\omega_p^2 + \omega_n^2}{j\dfrac{\omega_p^2}{Q_p}} \qquad (4.11)$$

Fig.(4.18) illustrates the "ideal" theoretical Notch filter amplitude response for a variety of frequencies and constant Qp(Q=2), while Fig.(4.19) for a variety of Qps and for constant p (fp=3.83KHZ).

Fig. (4.20) illustrates the effect of the frequency dependency of Ai, (i=1, 2) on the response.

5. <u>All Pass (AP) Realization</u>

As proposed in Table (3.1) using the same admittances values as Notch. An All Pass transfer function can be derived as

$$T_1(s) = \frac{s^2 - \dfrac{\omega_p}{Q_p}s + \omega_p^2}{s^2 + \dfrac{\omega_p}{Q_p}s + \omega_p^2} \qquad (4.12)$$

which takes the following values

$$\text{at } \omega \to 0, \quad |T_1(j0)| \to 1 = 0\,\text{db}, \quad \underline{/T_1(j0)} \to 360^\circ$$

$$\omega \to \infty, \quad |T_1(j\infty)| \to 2 = 6\,\text{db}, \quad \underline{/T_1(j\infty)} \to 0^\circ \tag{4.13}$$

The above agree with the computer simulation results of Figs. (4.21) and (4.22).

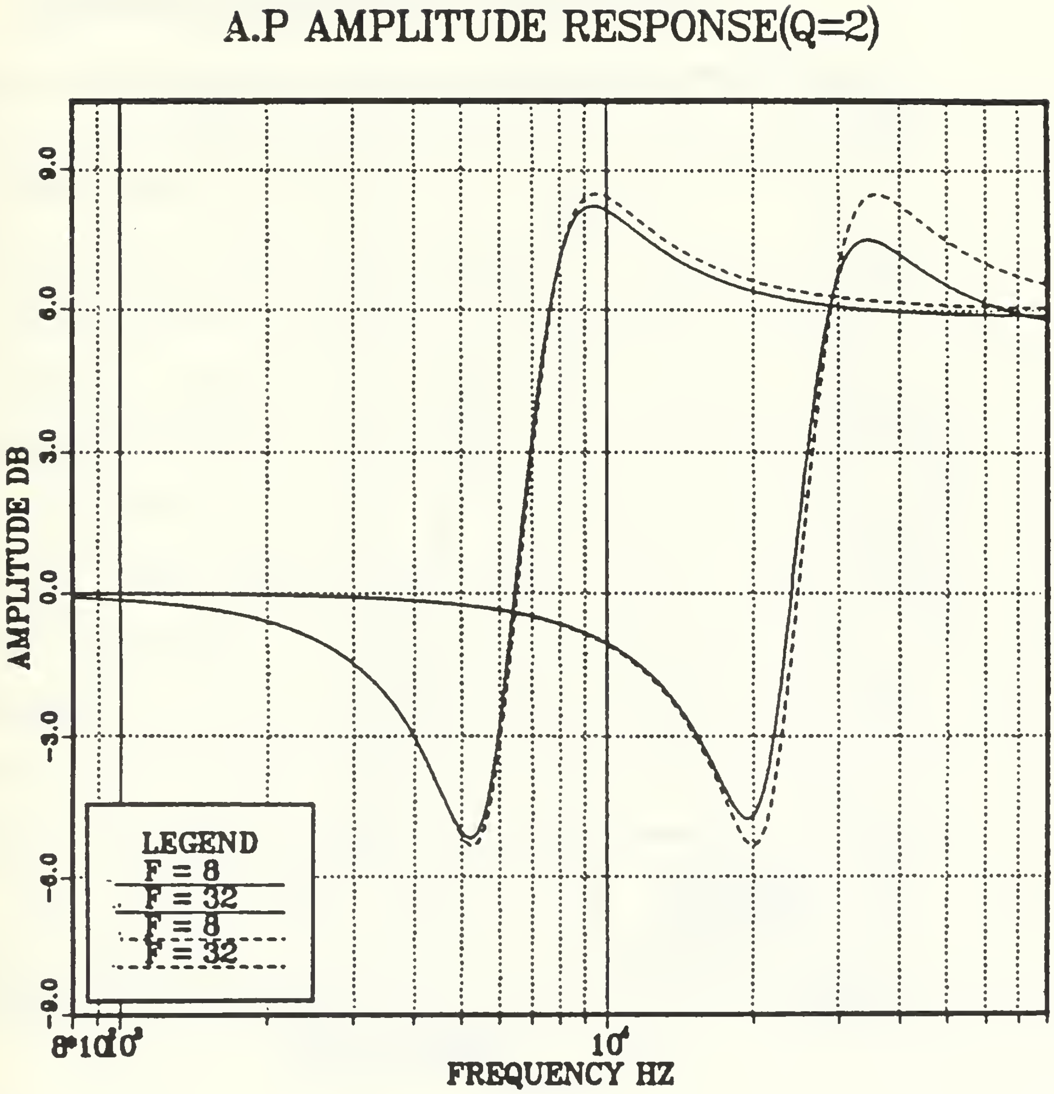


Fig. 4.21 - Ideal vs. Nonideal APF Amplitude Response.

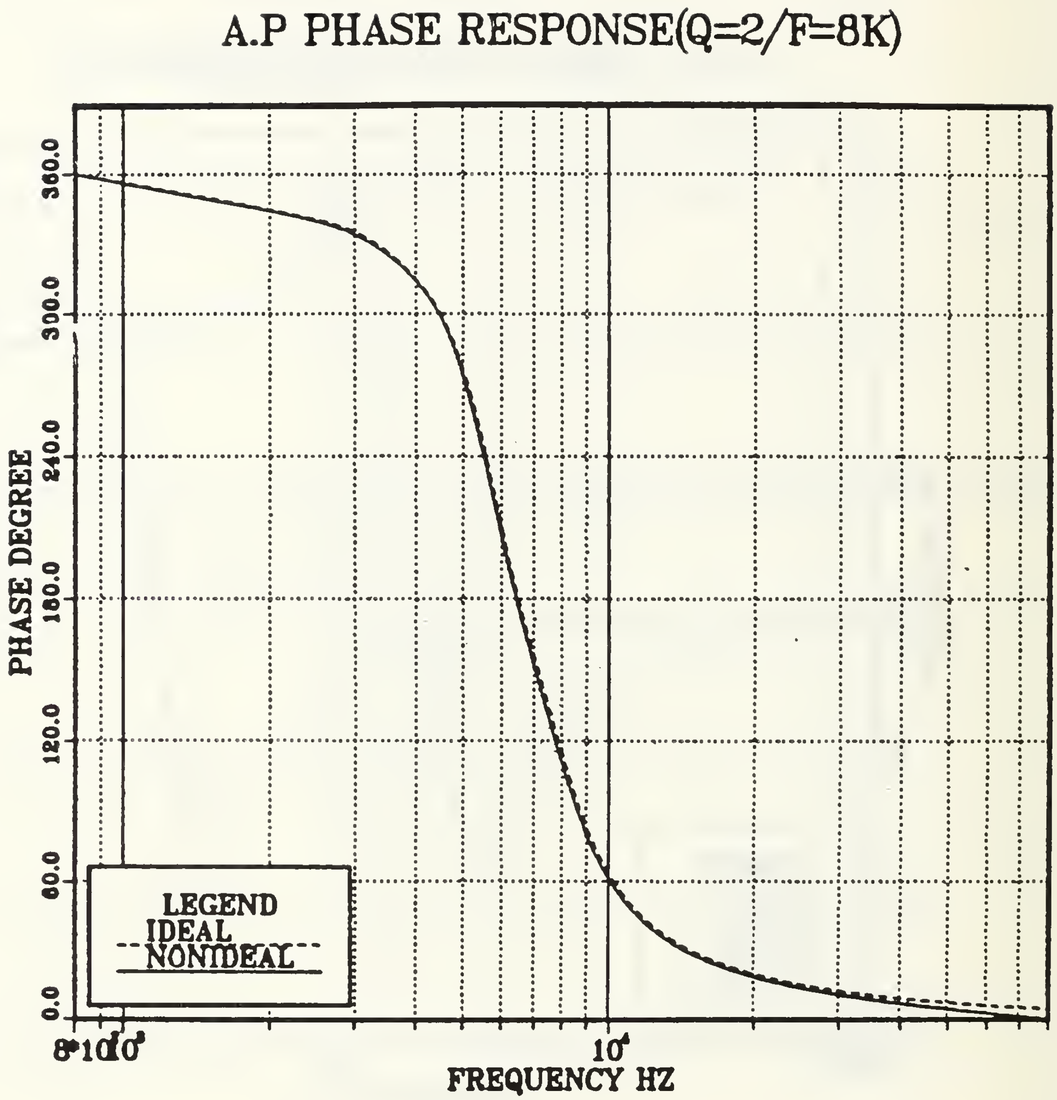


Fig. 4.22 - Ideal vs. Nonideal Phase Response.

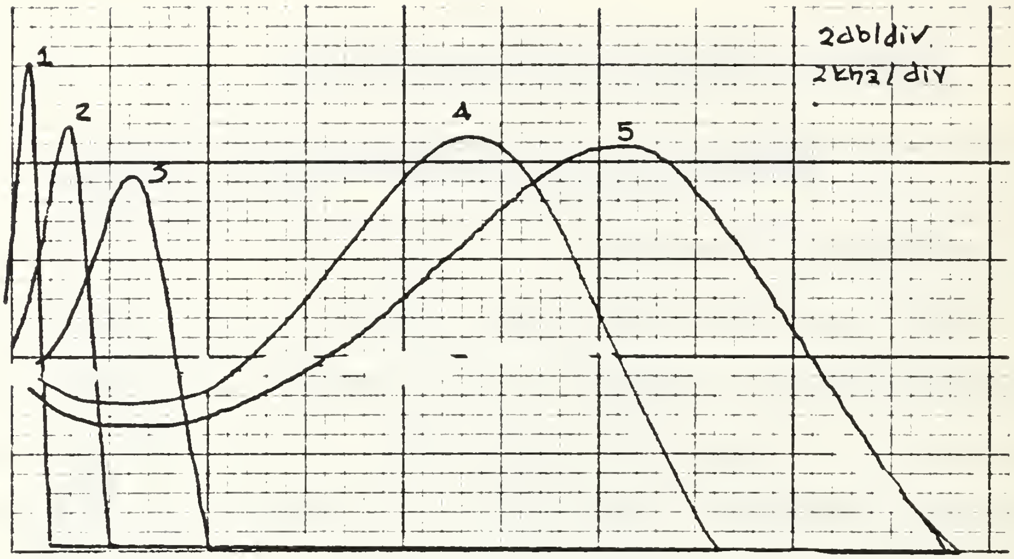
V. REALIZATION OF PROGRAMMABLE GIC FILTER

A. EXPERIMENTAL RESULTS

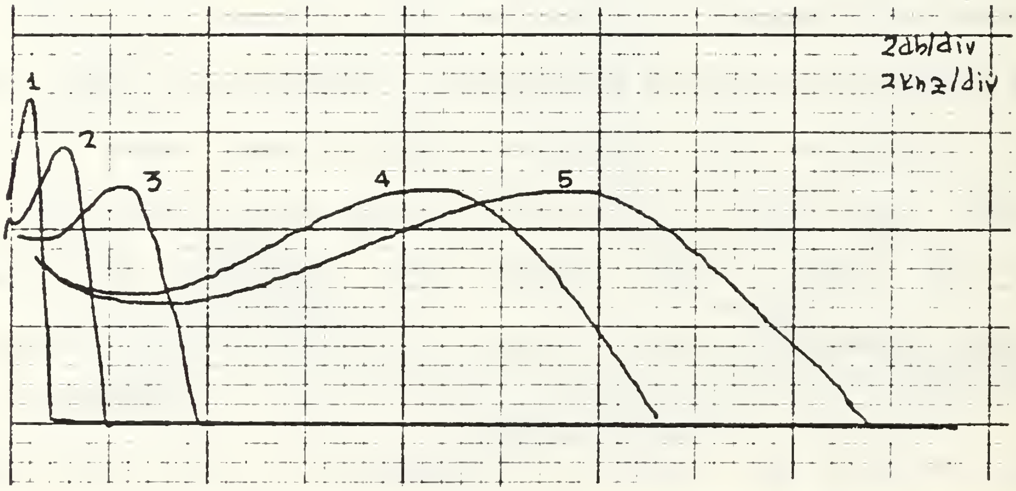
After the circuit of Figs. (3.10), (3.9), (3.8), (3.7), was constructed a variety of measurements were taken in order to study the response of the network to the different inputs (control bitwords). To observe the affect of the control switches which introduce a resistance of 80 each at CLOSED position, two values of Rs, Rq, and Cs were used with one decade difference in magnitude. That means that R was given the values of 1.6K (as discussed in Chapter III) and 16K, the four resistors that consisted the Rq bank were of values (1.6K, 3.2K, 6.4K, 12.8K) in the first case and (16K, 32K, 64K and 128K) in the second one, and that the capacitor bank's capacitors were chosen of values (100nF, 50nF, 12, 5nF, 11nF) and (10nF, 5nF, 1.2nF, 0.1nF) accordingly, to be able to keep the range of frequencies as much the same as possible for both cases.

1. Low Pass Filter

With the topology-control bit word 000 the network realized a LPF response. Fig. 5.1(a) illustrates the LP response for R=1.6KΩ to a variety of frequencies for Q=5 while 5.1(b) illustrates the same but for Q=2. It can be observed from 5.1(a) (Q=5) that the amplitude response



(a)



(b)

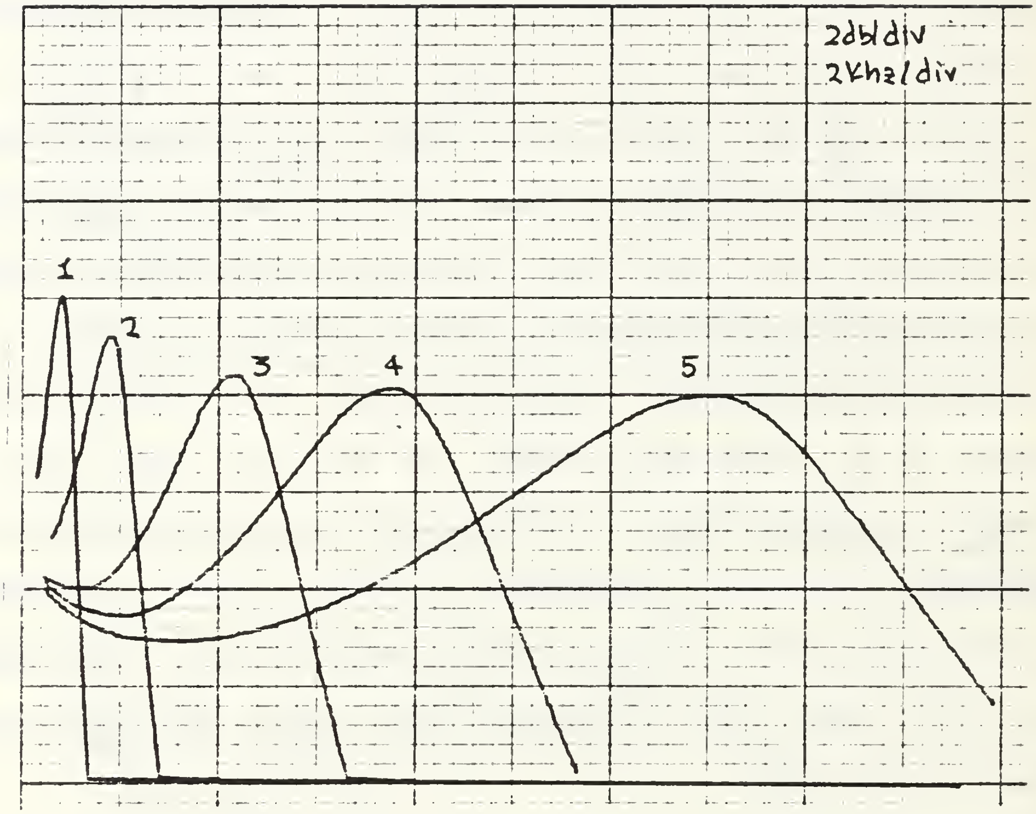
1 - f = 0.99khz
2 - f = 1.98khz
3 - f = 3.83khz
4 - f = 7.96khz
5 - f = 12.0khz

Fig. 5.1 - (a) LPF Response for R=1.6k (Q=5, Rq=8.0k)
(b) LPF Response for R=1.6k (Q=2, Rq=3.2k)

decreases while the frequency increases as it was expected from computer simulation. This is occurred up to the frequency of 6KHZ; then it started increasing with the frequency, while at Fig. (5.1(b)) (Q=2) it remained constant. Fig. (5.2) illustrates (for R=16KHZ) that the observation in Fig (5.1) is not any more the case (for that frequency range) and the network responses the same as in computer simulation while in Fig. (5.3) (Q=5) the above can be noticed again. this is due to the interference of the control switches as it was discussed previously. Figs. (5.4) and (5.5) illustrate a variation of Q values for f=9KHZ for the two cases (R=1.6k and R=16K) respectively. A difference in magnitude can be observed due to the interference of the control switches. Figs. (5.6) and (5.7) illustrate the same but for F=12.8KHZ. Figs. (5.8) and (5.9) illustrate the phase and amplitude response of the LPf.

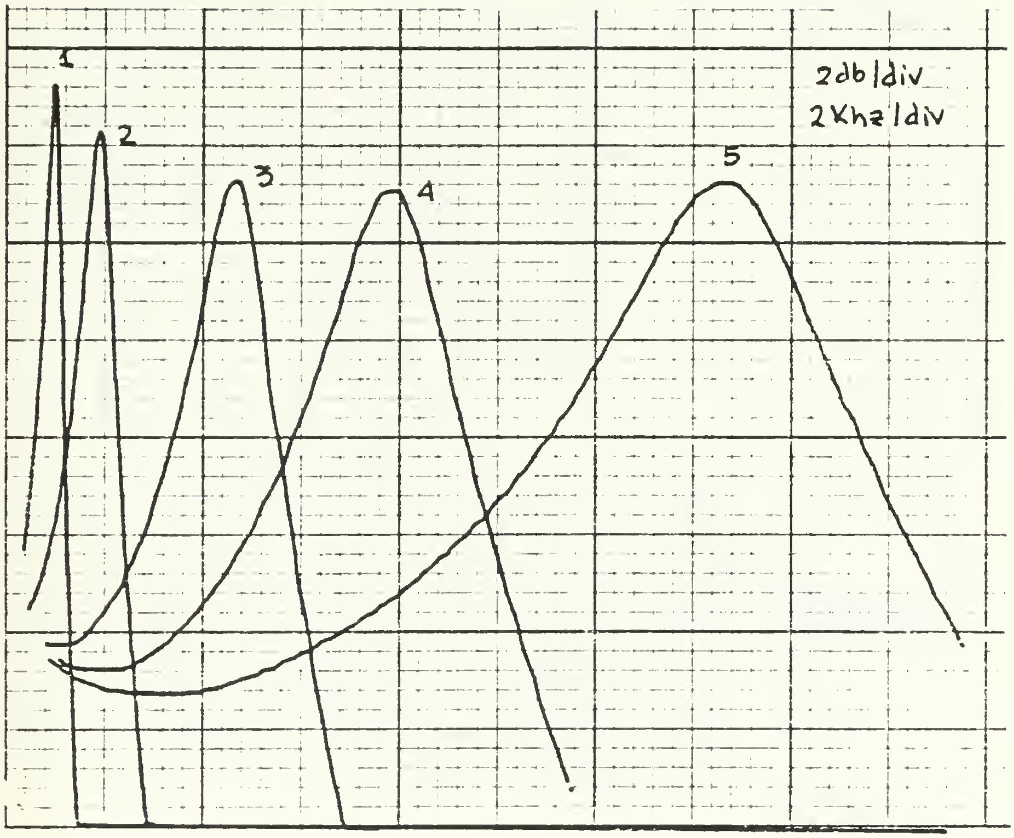
2. High Pass (HP) Realization

With the topology-control bitword 001, the network realized a high pass filter. Figs. (5.10) and (5.11) illustrate the HPF amplitude response for a variety of frequencies and (Q=2). It can be observed (as in LPF realization) that for this frequency range and for R=1.6KΩ the amplitude starts increasing as the frequency increases, decreases then again while for R=16KΩ this does not occur. the same observation submerges comparing Fig.(5.14) and (5.15). Figs. (5.12) and (5.13) are the plots obtained for



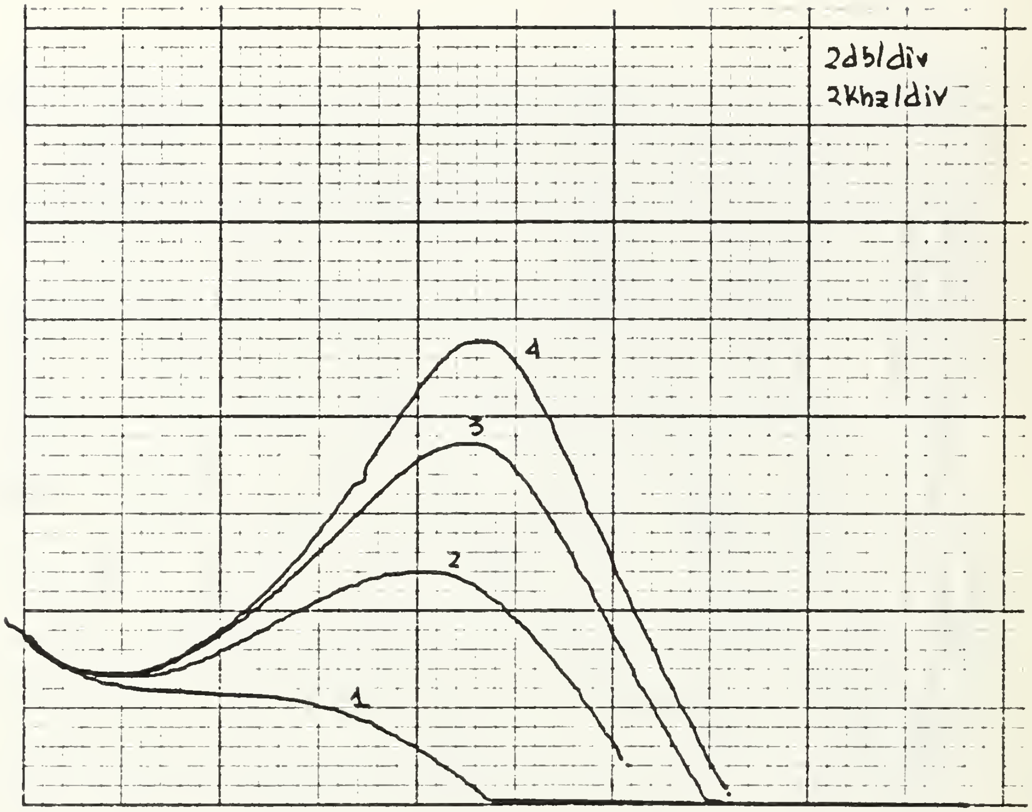
1 - f = 0.99khz
2 - f = 1.99khz
3 - f = 3.83khz
4 - f = 7.90khz
5 - f = 12.8khz

Fig. 5.2 - LPF Amplitude Response
R=1.6k (Q=2, Rq=32k)



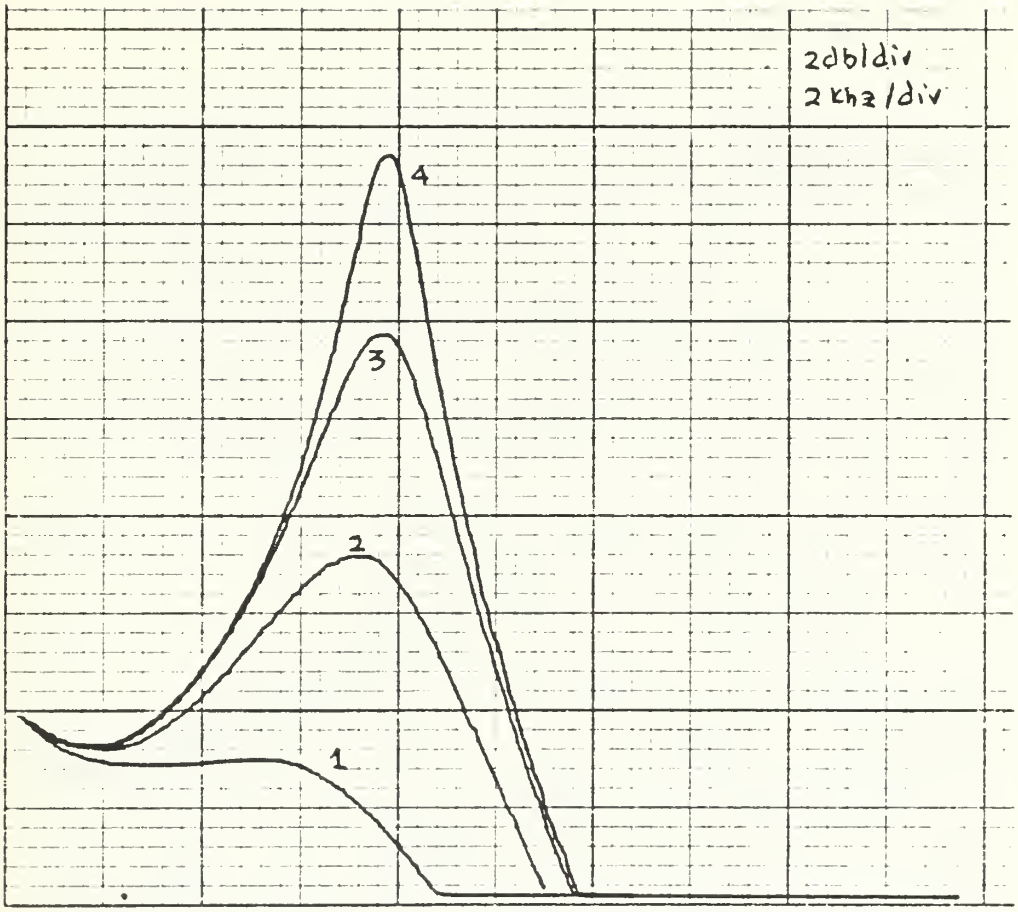
1 - f = 0.99khz
2 - f = 1.99khz
3 - f = 9.20khz
4 - f = 7.96khz
5 - f = 15.0khz

Fig. 5.3 - LPF Amplitude Response
R=16k (Q=5, Rq=50k)



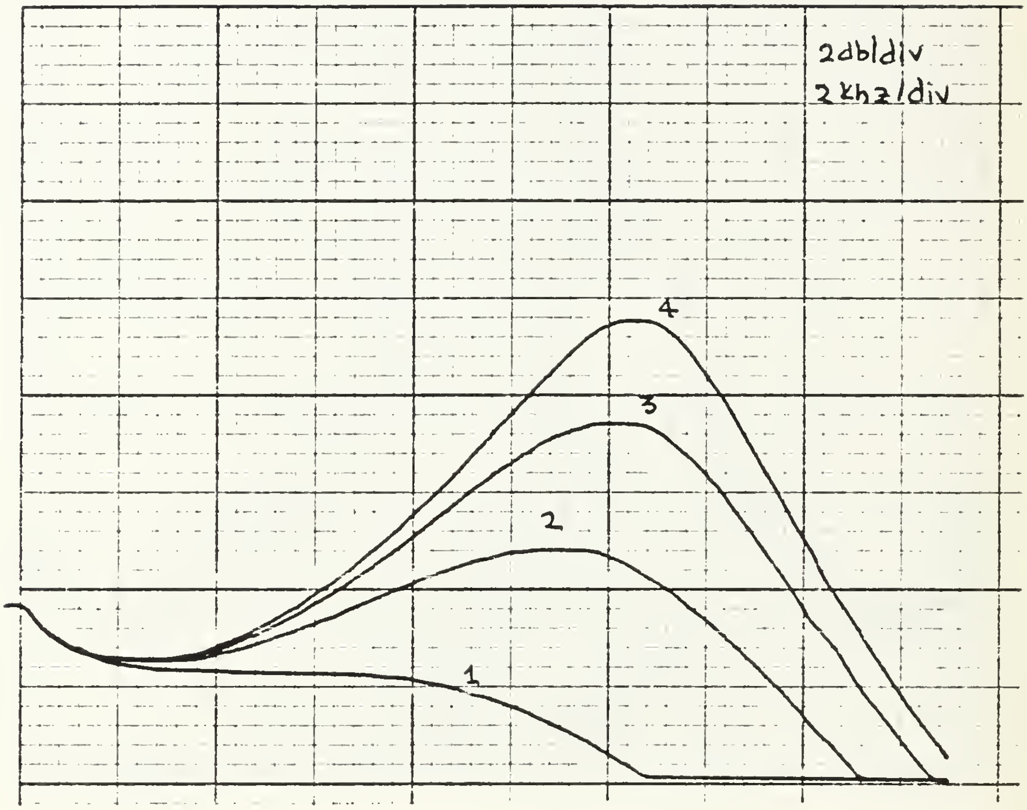
1 - Q = 1
2 - Q = 2
3 - Q = 4
4 - Q = 7

Fig. 5.4 - LPF Amplitude Response (R=1.6k),
f=7.96khz



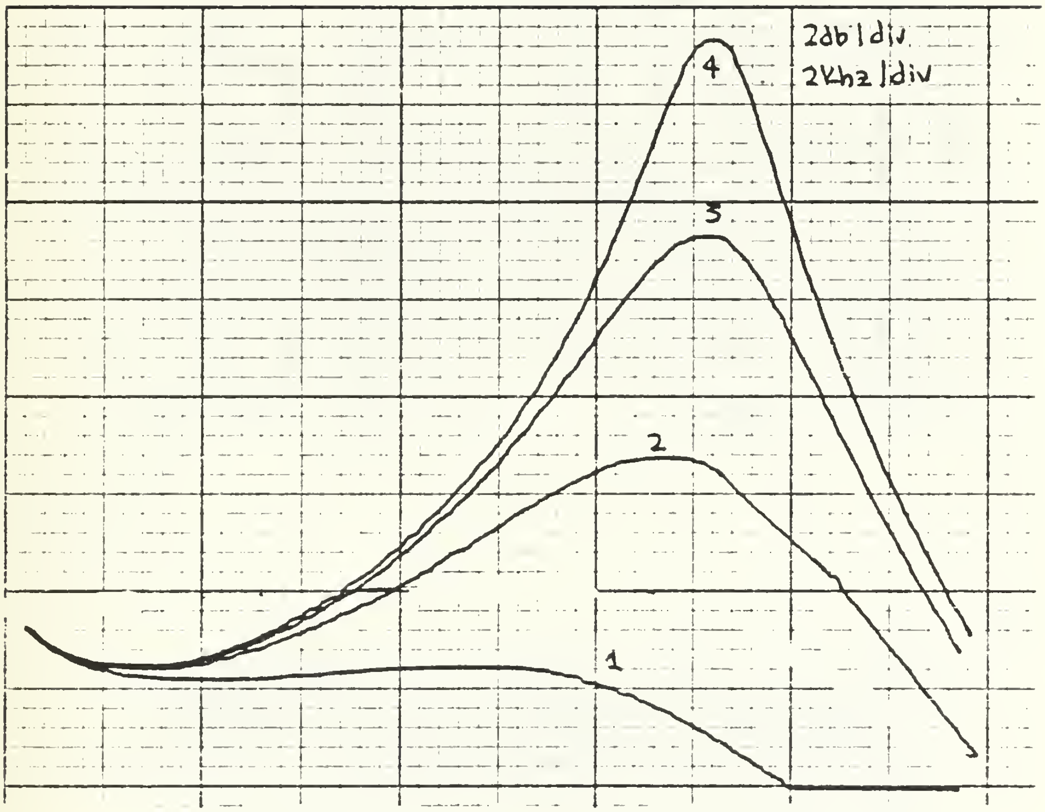
1 - Q = 1
2 - Q = 2
3 - Q = 4
4 - Q = 7

Fig. 5.5 - LPF Amplitude Response (R=10k), f=7.96khz



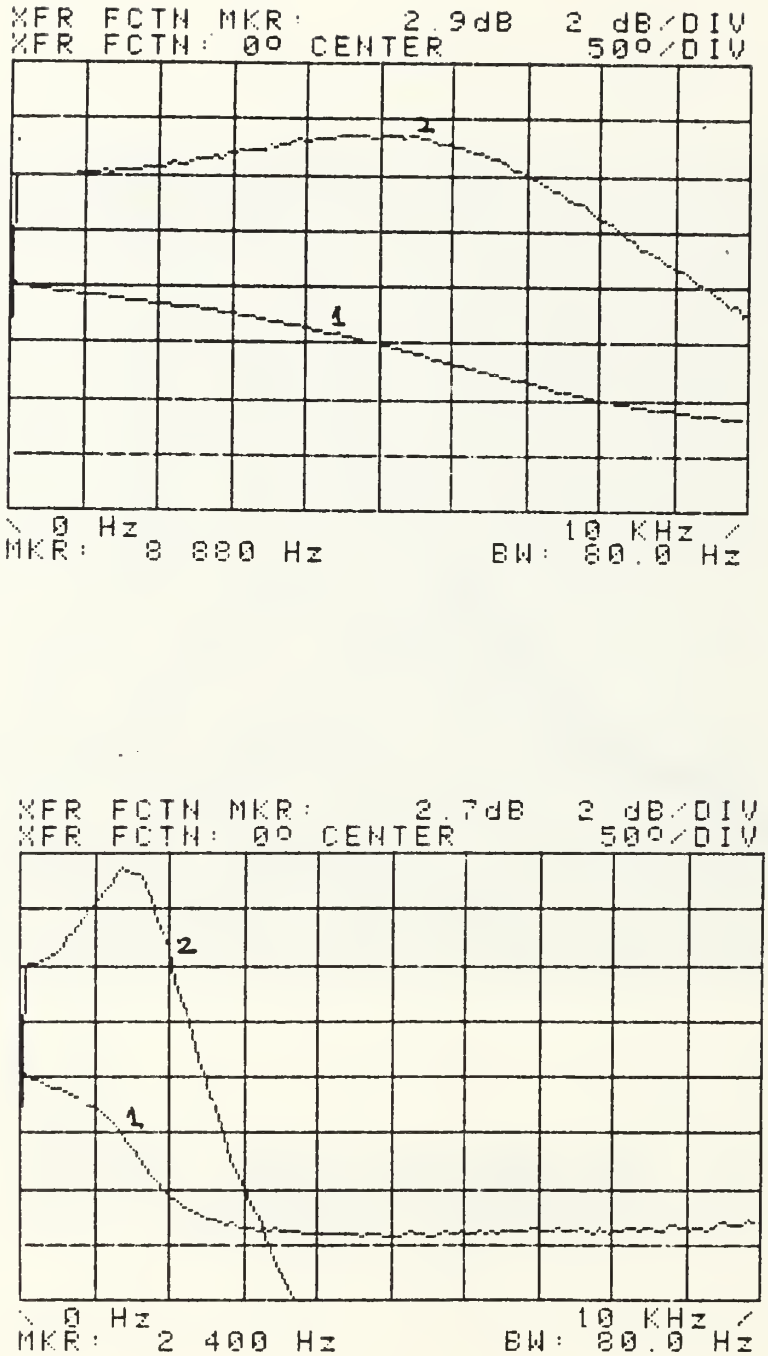
1 - Q = 1
2 - Q = 2
3 - Q = 4
4 - Q = 7

Fig. 5.6 LPF Amplitude Response (R=1.6k) and f=12.8khz



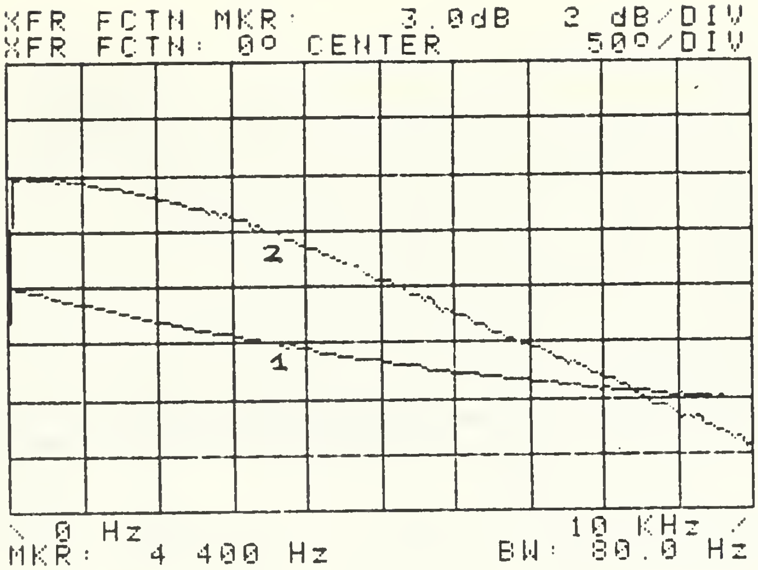
1 - Q = 1
2 - Q = 2
3 - Q = 4
4 - Q = 7

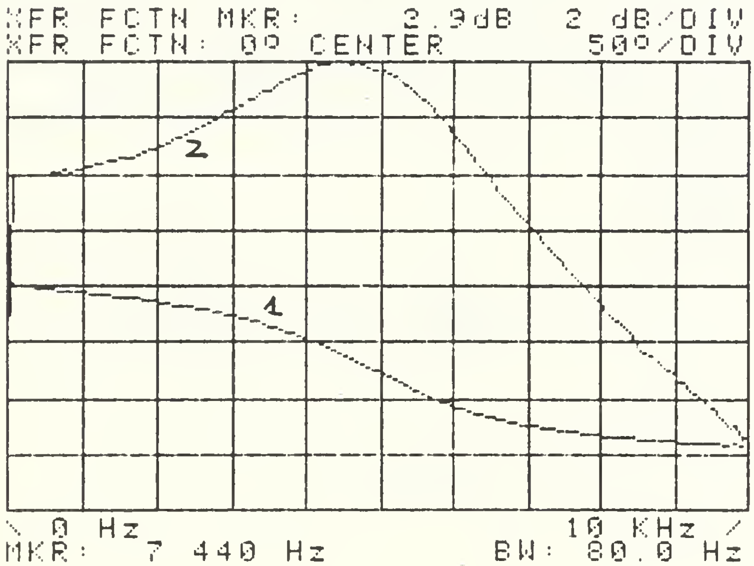
Fig. 5.7 - LPF Amplitude Response (R=1.6k), f=12.8khz



1. Phase Response.
2. Amplitude Response.

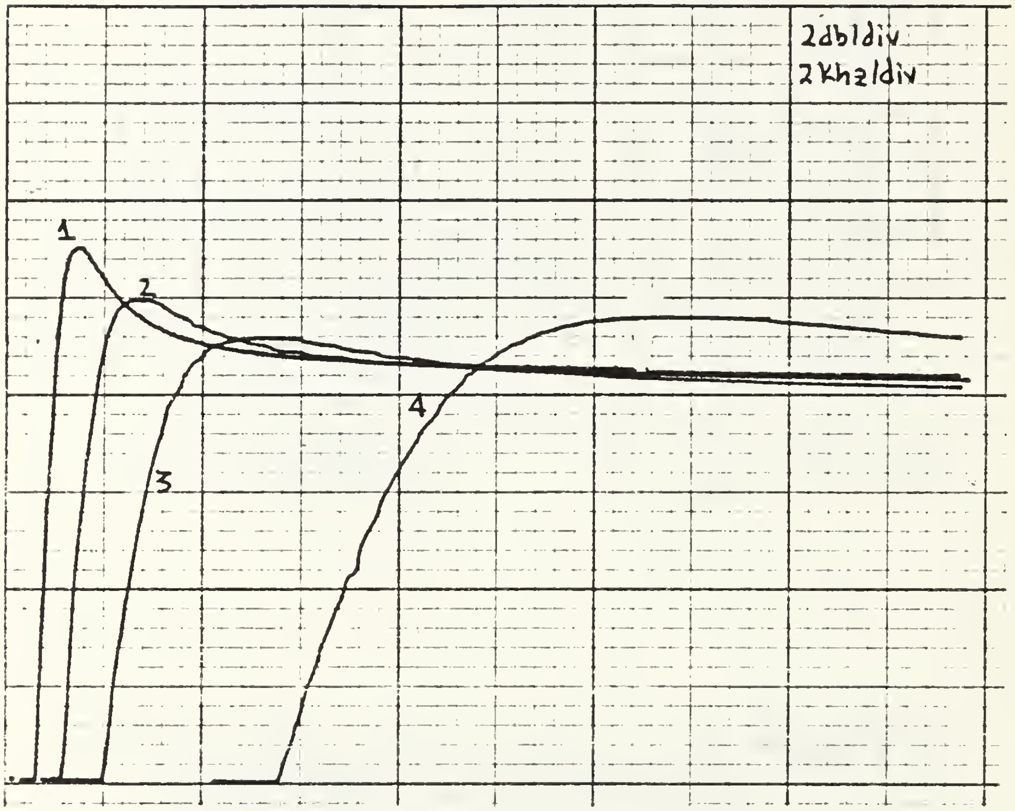
Fig. 5.8 - LPF Amplitude/Phase Response.





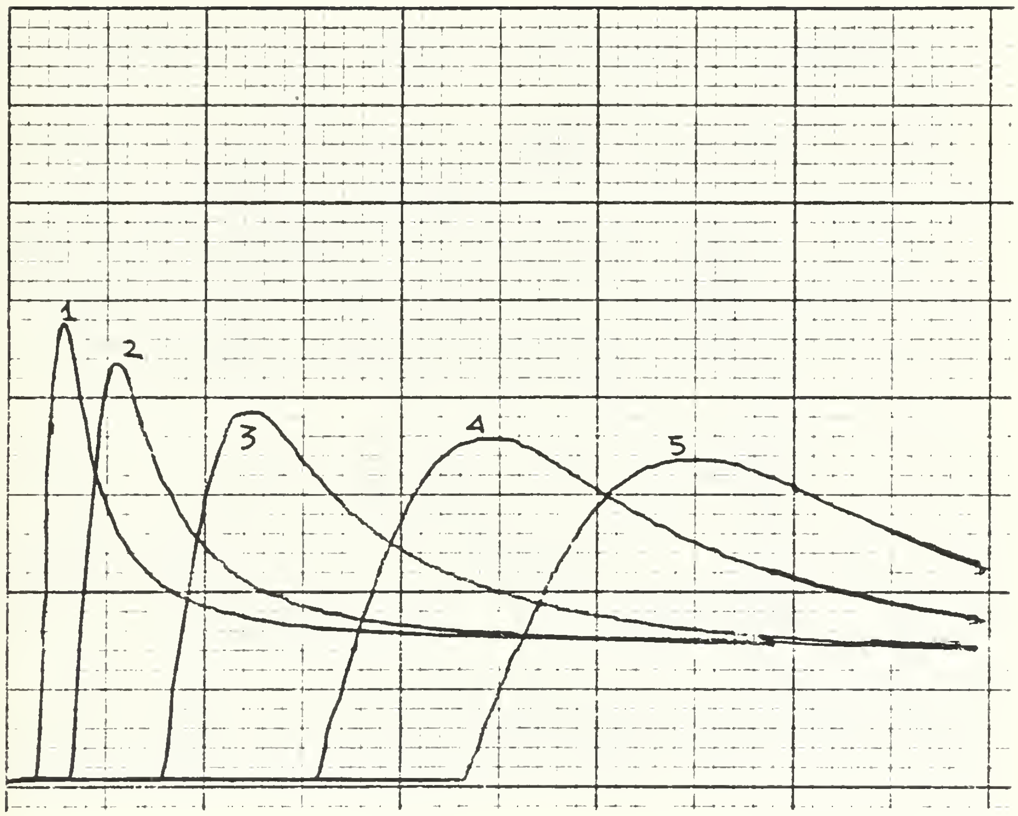
1. Phase
2. Amplitude

Fig. 5.9 - LPI Phase/Amplitude Response.



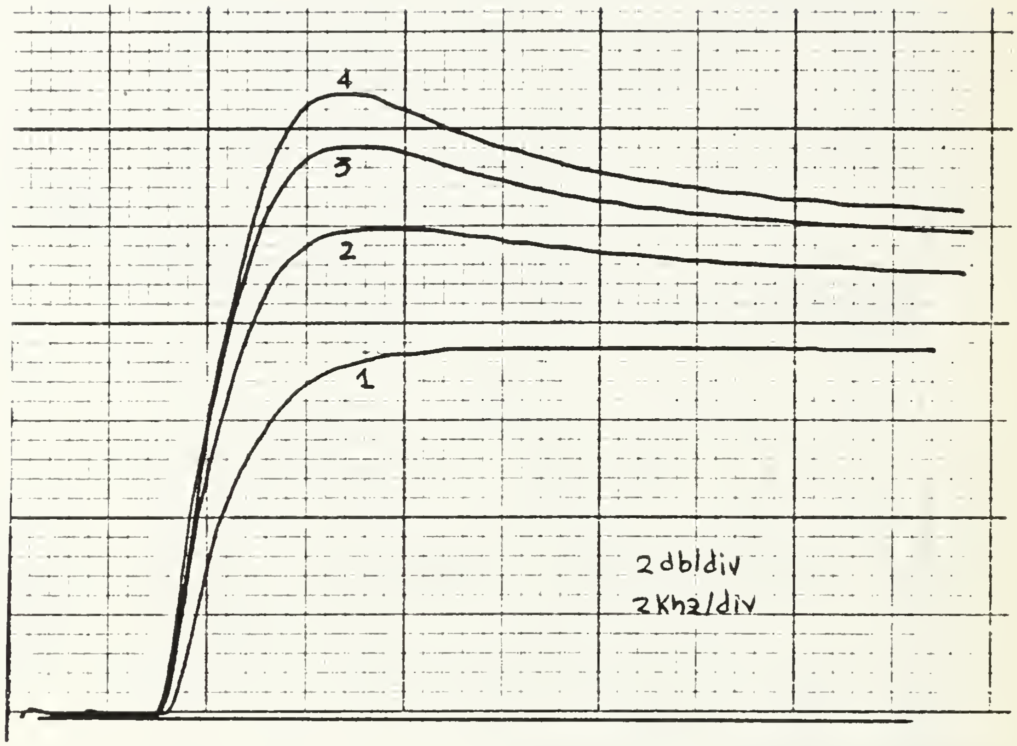
1 - f = 0.99khz
2 - f = 1.99khz
3 - f = 3.83khz
4 - f = 7.96khz

Fig. 5.10 - HPF Amplitude Response
(R=1.6k), Q=2



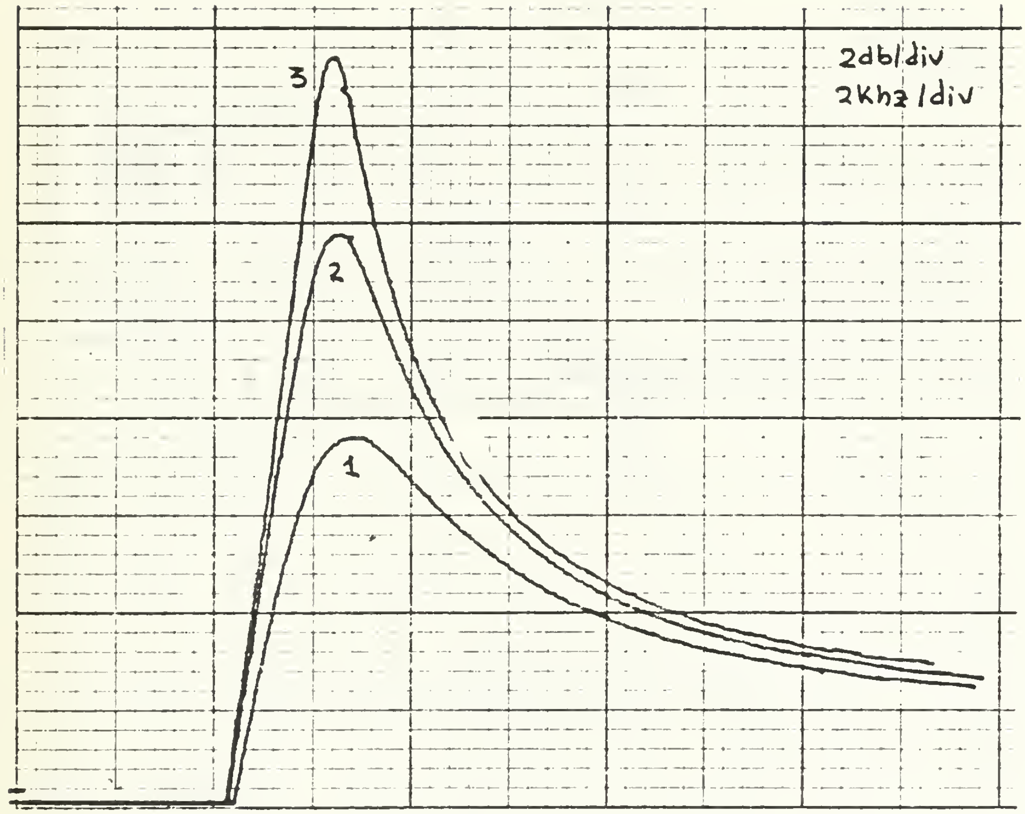
1 - f = 0.99khz
2 - f = 1.9 khz
3 - f = 5.0 khz
4 - f = 8.0 khz
5 - f = 13.8 khz

Fig. 5.11 - HPF Amplitude Response
(R=1.6k), Q=2



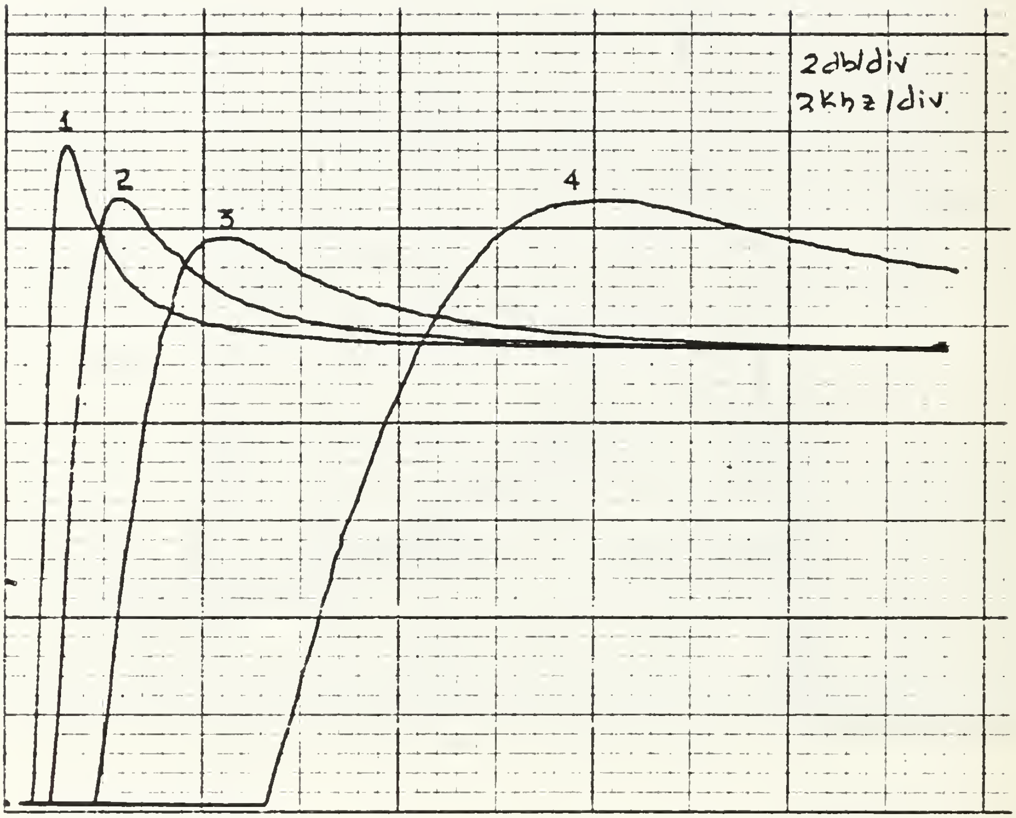
1 - Q = 0.7
2 - Q = 2.0
3 - Q = 4.0
4 - Q = 7.0

Fig. 5.12 - HPF Response (R=1.6k),
f=6.65khz



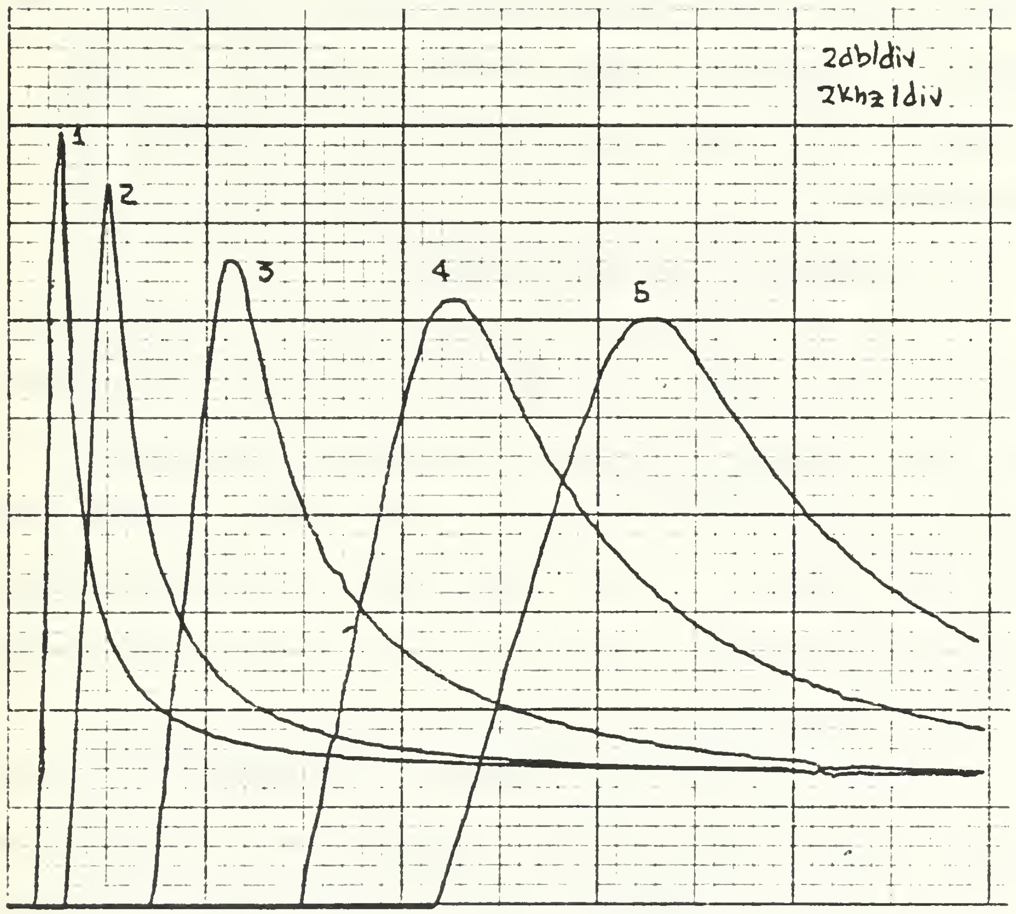
1 - Q = 2.0
2 - Q = 4.0
3 - Q = 7.0

Fig. 5.13 - HPF Amplitude Response (R=1.6k), f=6.65khz



1 - f = 0.99khz
2 - f = 1.99khz
3 - f = 3.83khz
4 - f = 7.96khz

Fig. 5.14 - HPB Amplitude Response
(R=1.6k), Q=5



1 - f = 0.99khz
2 - f = 1.99khz
3 - f = 3.83khz
4 - f = 7.96khz
5 - f = 12.8khz

Fig. 5.15 - HPB Amplitude Response
(R=1.6k), Q=5

f=6.65KHZ and a variety of Qs for R=1.6KΩ and R=16KΩ , respectively. a difference of about 12dB in amplitude appears. Finally, figs (5.16), (5.17), (5.18), (5.19) illustrate the phase and amplitude responses of the HP realization.

3. <u>Band Pass (BP) Realization</u>

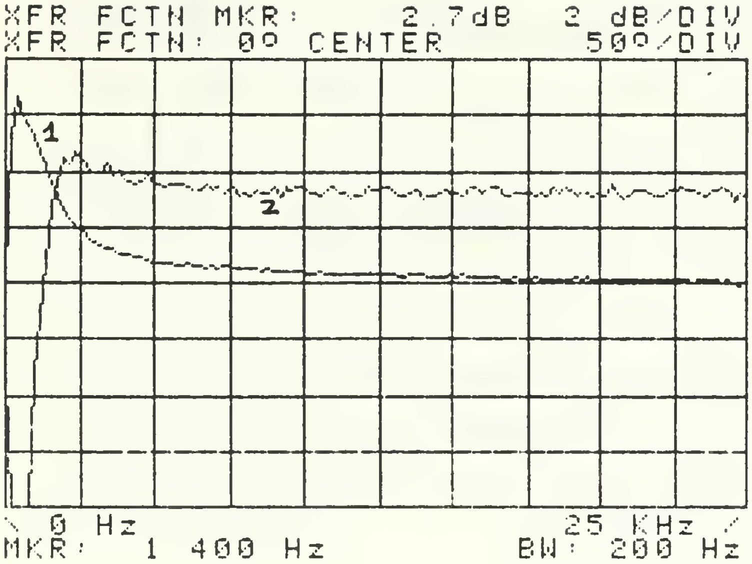
The 010 topology-control bitword realizes the Band Pass Filter. Figs. (5.20) and (5.21) illustrates the amplitude response for set of frequencies and Q=10. At both cases the amplitude decreases while the frequency increases until the value of 9KHZ. Then starts increasing again. It can be also observed that for R=1.6K the frequency deviation from the theoretical fp is larger.

Figs. (5.22) and (5.23) illustrate the amplitude response again, but for Q=1. This time the deviations from the theoretical response (both of amplitude fluctuations and frequency shift) are less.

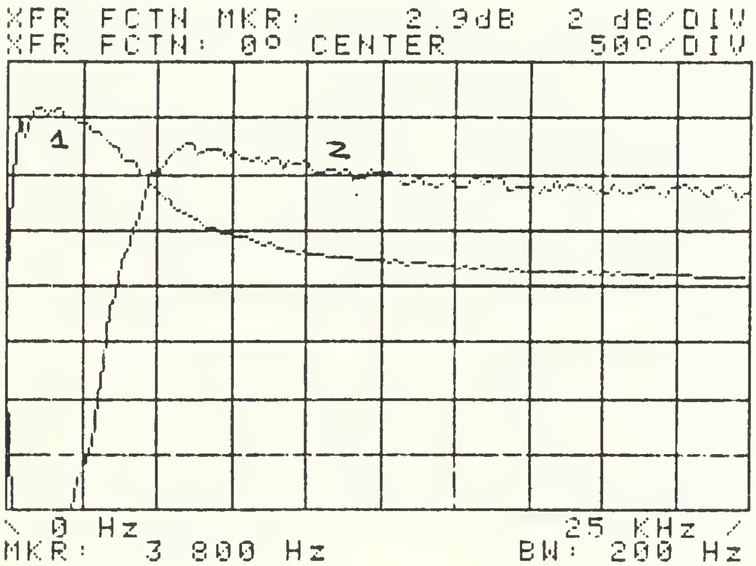
Figs. (5.24), (5.25), (5.26), and (5.27) are also plots of the amplitude response but for variation of Q. A difference of approximately 1dB appears for the lowest frequency (3.8KHZ) and of 0.5dB for the higher (9KHZ), Finally, Figs. (5.28)-(5.32) illustrate the phase and amplitude of the BPF response.

4. <u>Notch (N) Filter Realization</u>

With the topology-control bitword 011, a Notch filter realization can be achieved. Figs. (5.33) and (5.34)



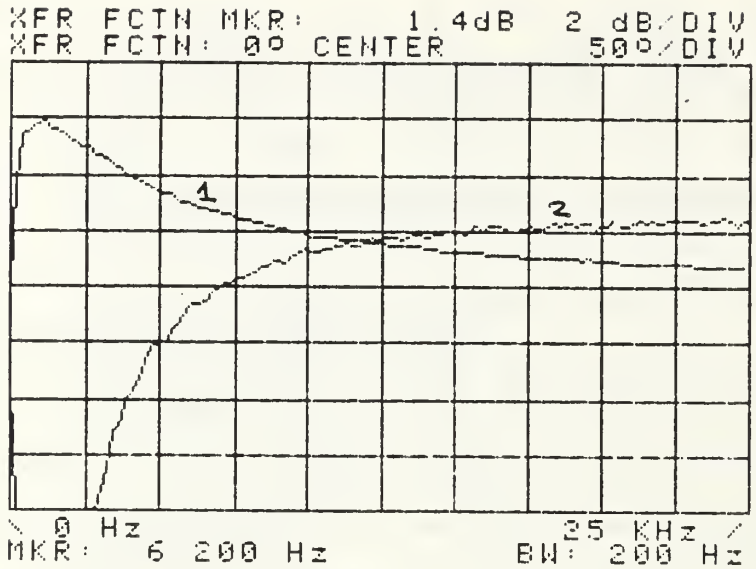
(a)

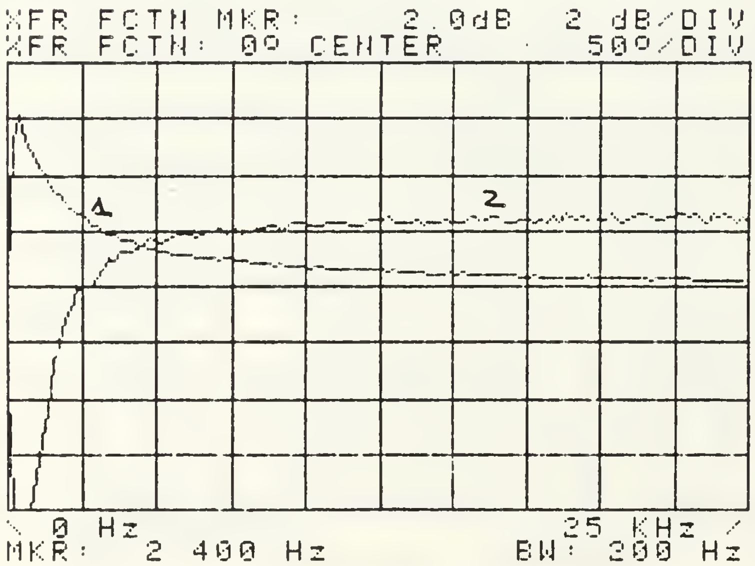


(b)

1. Phase
2. Amplitude

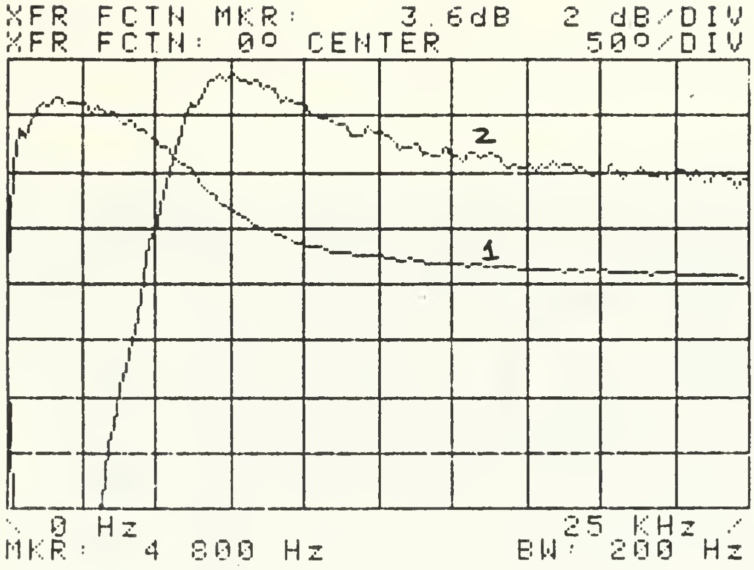
Fig. 5.16 - HPF Amplitude/Phase Response (Q=1)

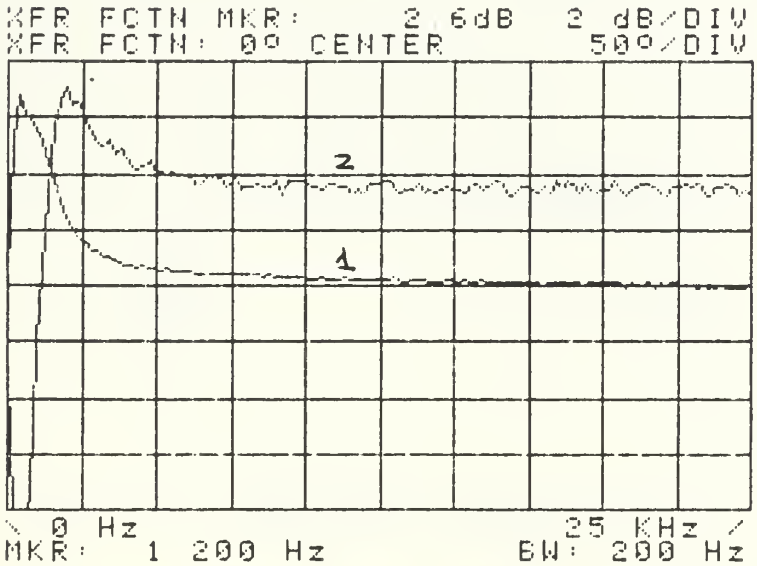




1. Phase
2. Amplitude

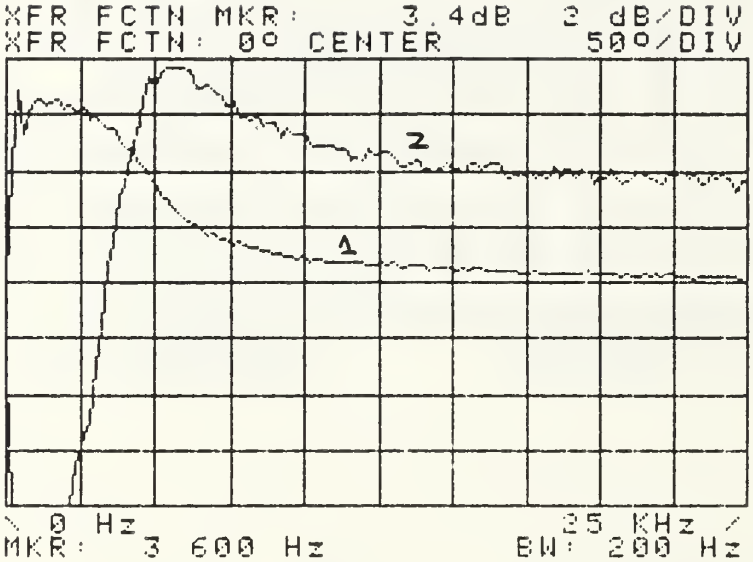
Fig. 5.17 - HBF Amplitude/Phase Response (Q=0.5)





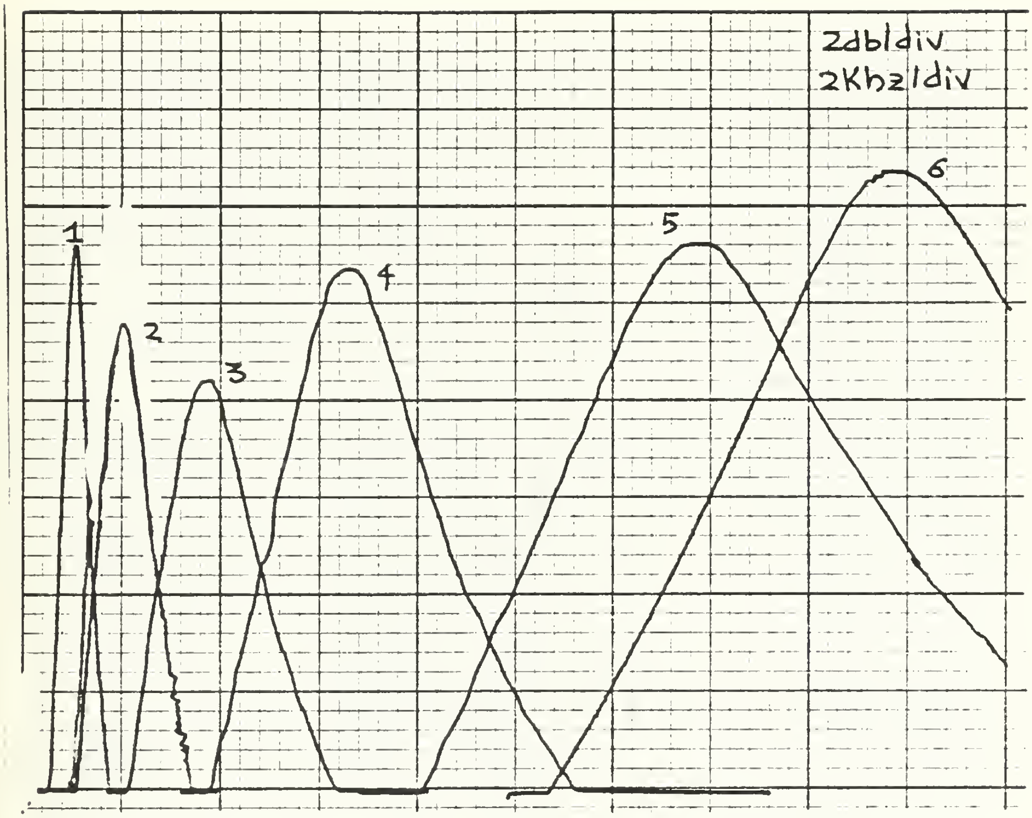
1. Phase
2. Amplitude

Fig. 5.18 - HPF Phase/Amplitude Response
(Q=2)



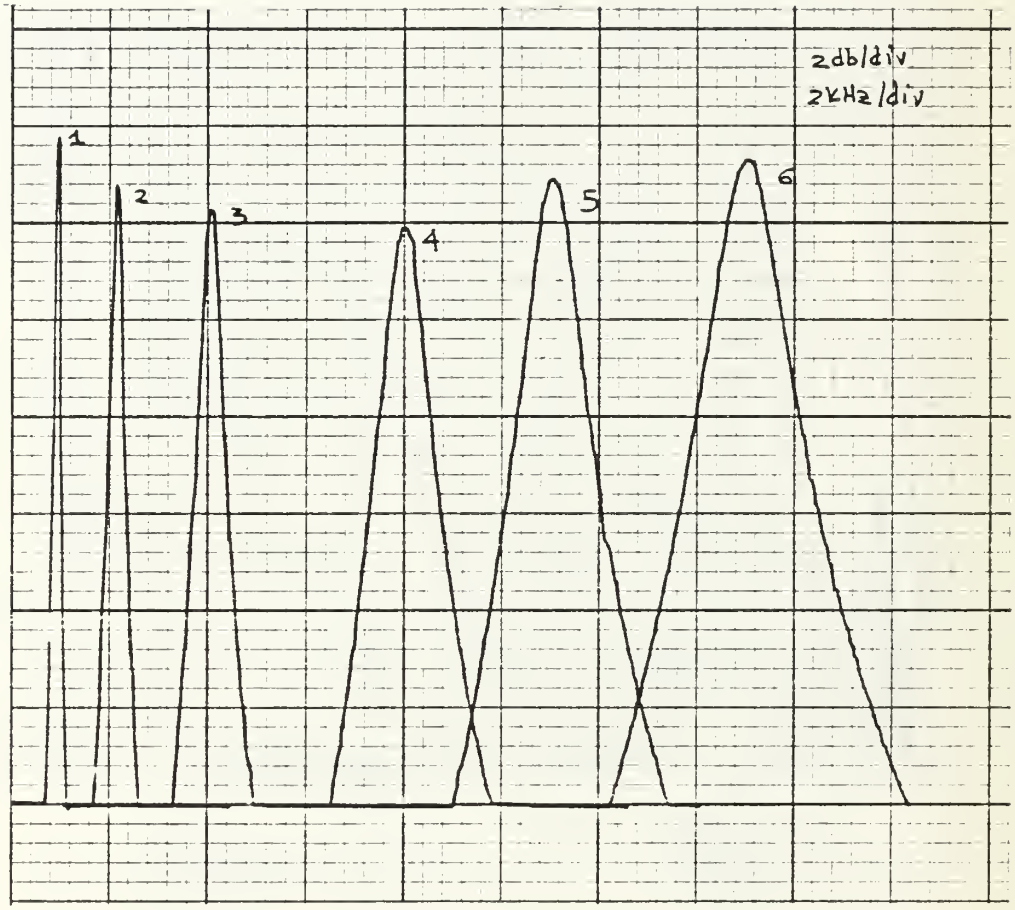
1. Phase
2. Amplitude

Fig. 5.19 - HPF Amplitude Response (Q=2)



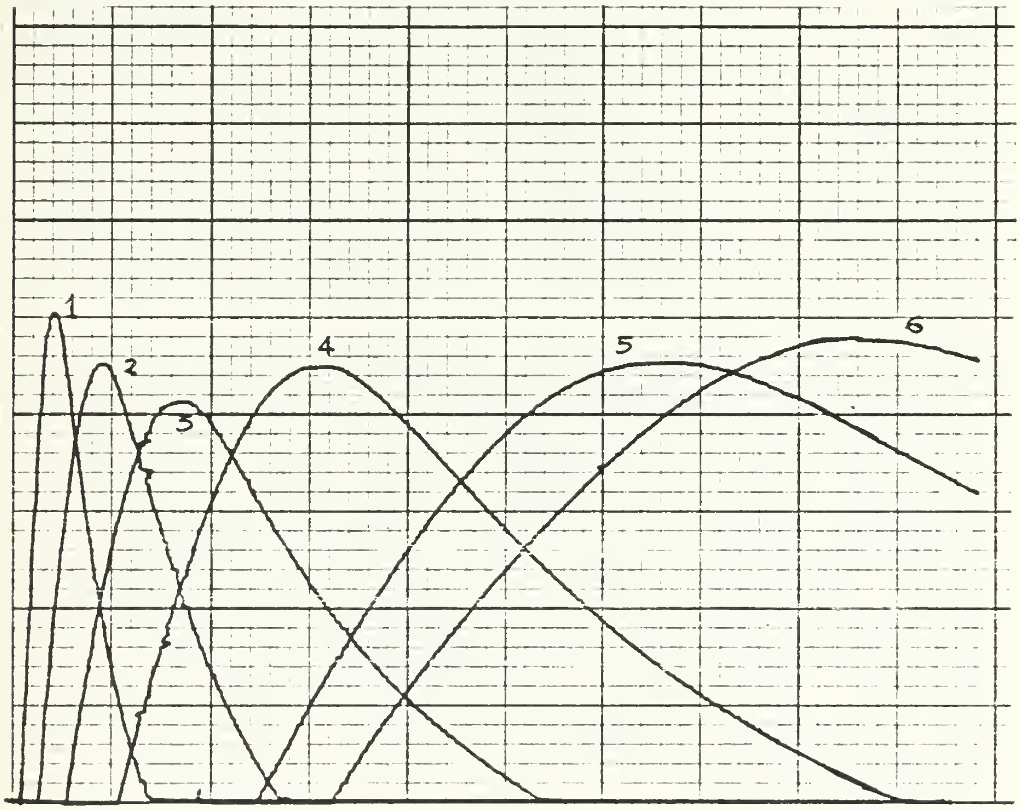
1 - f = 0.99khz
2 - f = 1.99khz
3 - f = 3.8khz
4 - f = 6.65khz
5 - f = 10.5khz
6 - f = 15.1khz

Fig. 5.20 - BPF Amplitude Response
(R=1.6k), Q=10



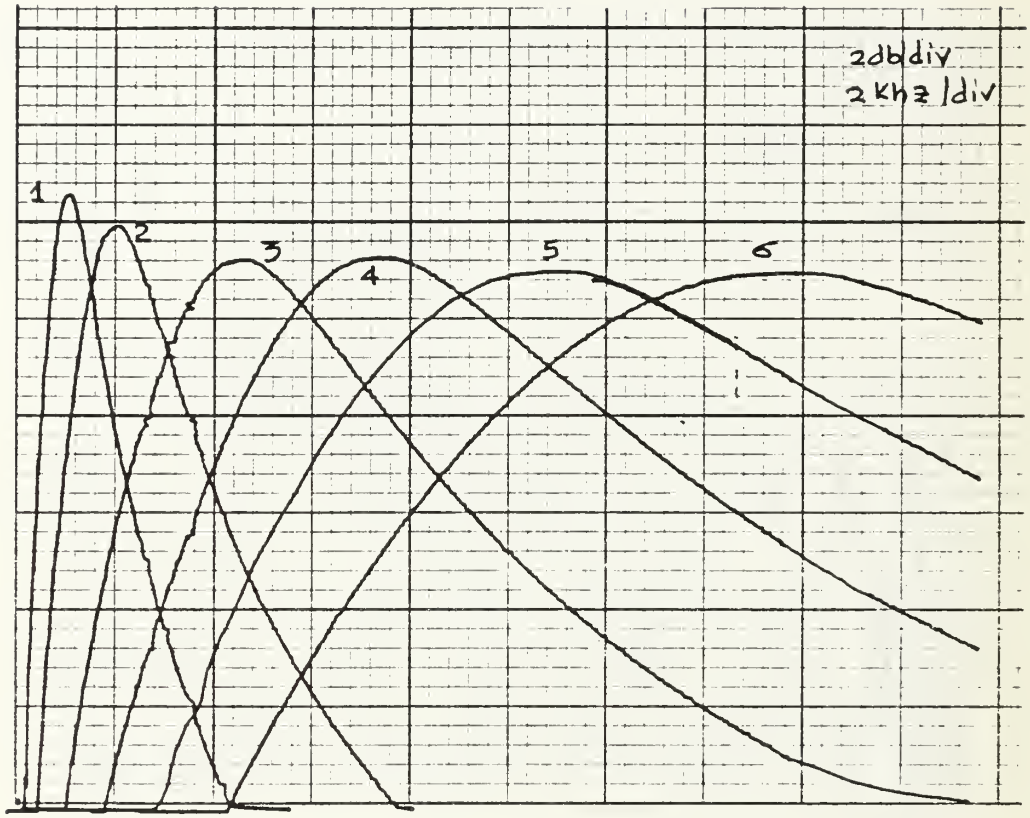
1. f = 0.99K
2. f = 1.99K
3. f = 3.8K
4. f = 7.96K
5. f =11.2 K
6. f =15.1 K

Fig. 5.21 - BPF Amplitude Response
(R=1.6k), Q=10



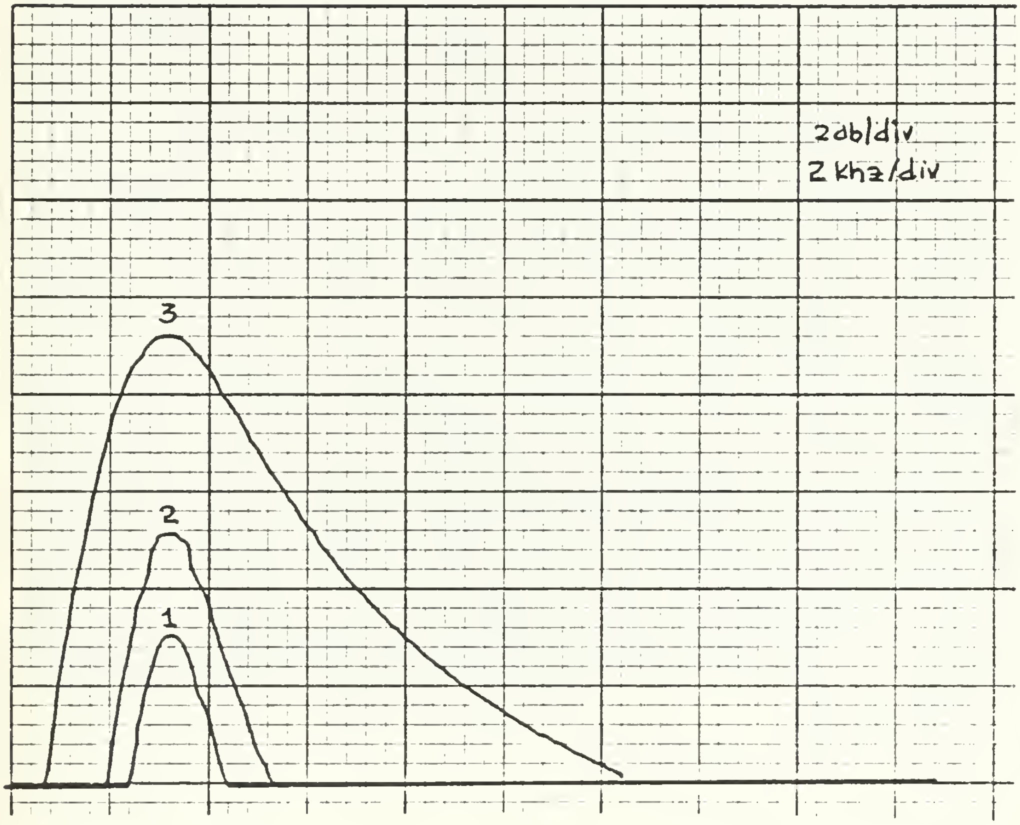
1. 0.99khz
2. 1.99khz
3. 3.8 khz
4. 6.65khz
5. 12.8khz
6. 15.1khz

Fig. 5.22 - BPF Amplitude Response
(R=1.6k), Q=7



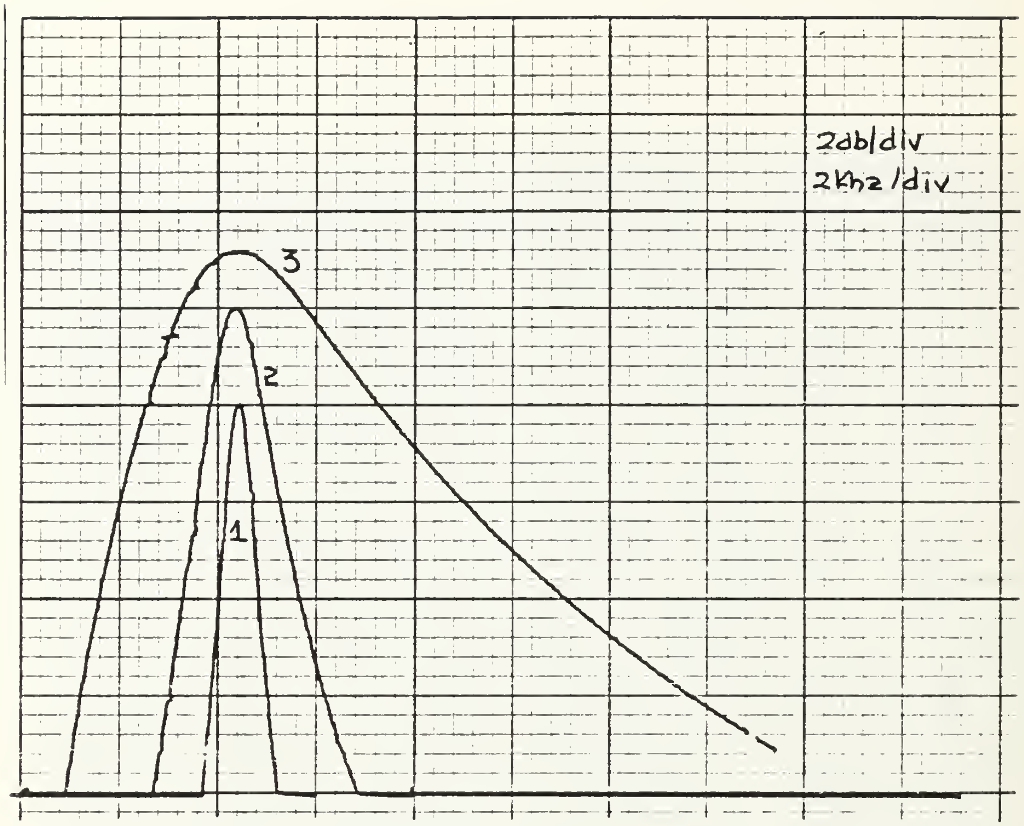
1. 0.99khz
2. 1.99khz
3. 3.83khz
4. 7.96khz
5. 12.8khz
6. 15. khz

Fig. 5.23 - BPF Amplitude Response (R=1.6k), Q=1



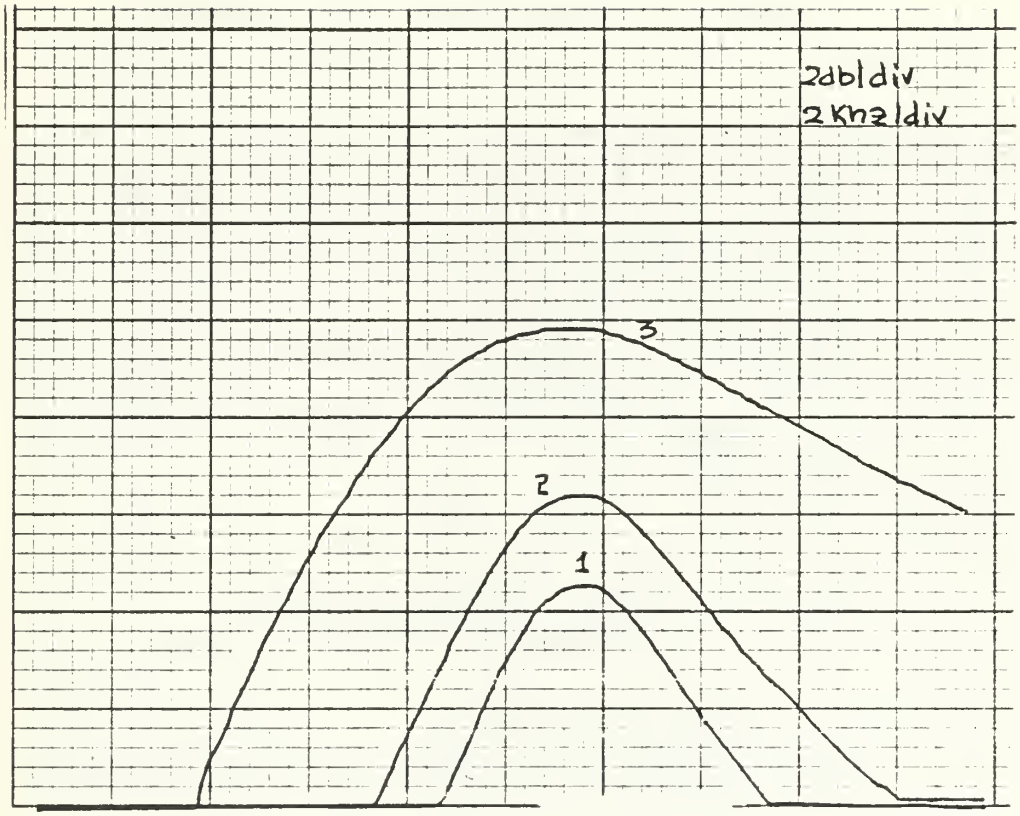
1 - Q = 10.0
2 - Q = 4.0
3 - Q = 1.0

Fig. 5.24 - BPI Amplitude Response (R=1.6k), f=3.83khz



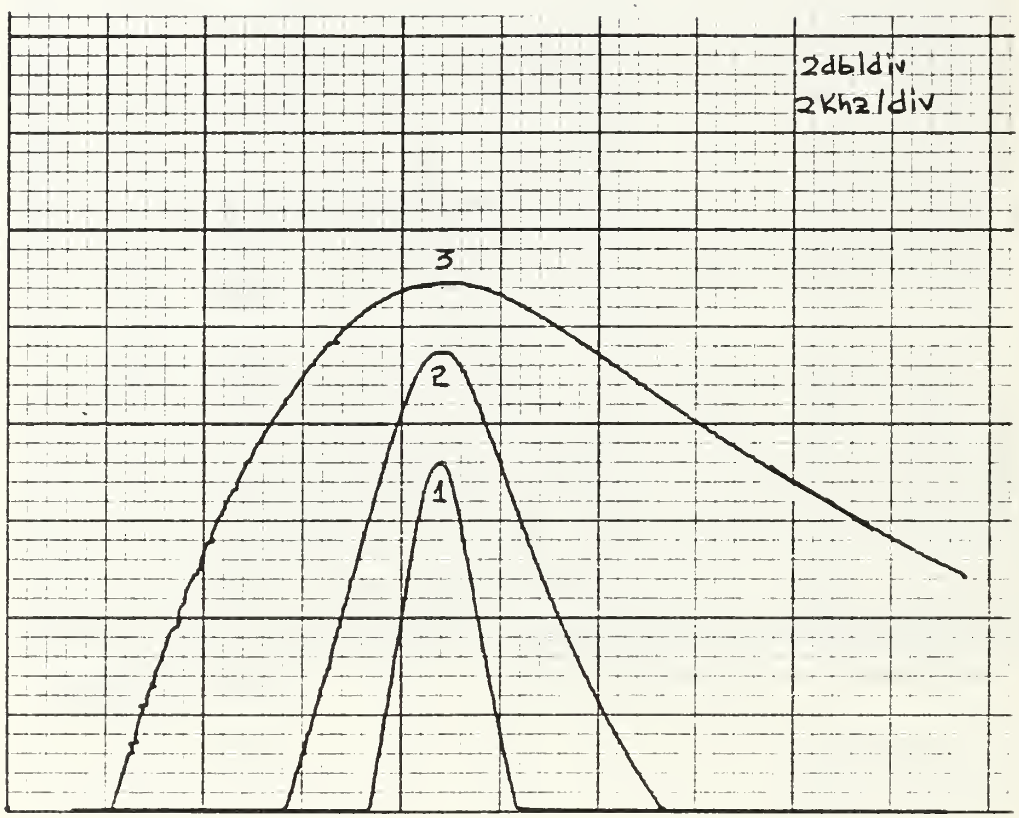
1 - Q = 10.0
2 - Q = 4.0
3 - Q = 1.0

Fig. 5.25 - BPI Amplitude Variation
(R=1.6k), f=3.83khz



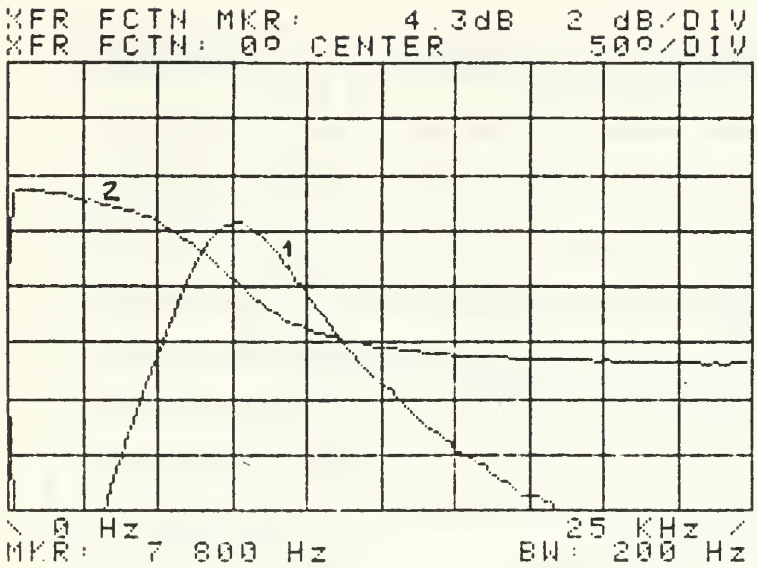
1 - Q = 10.0
2 - Q = 4.0
3 - Q = 1.0

Fig. 5.26 - BPF Amplitude Response
(R=1.6k)



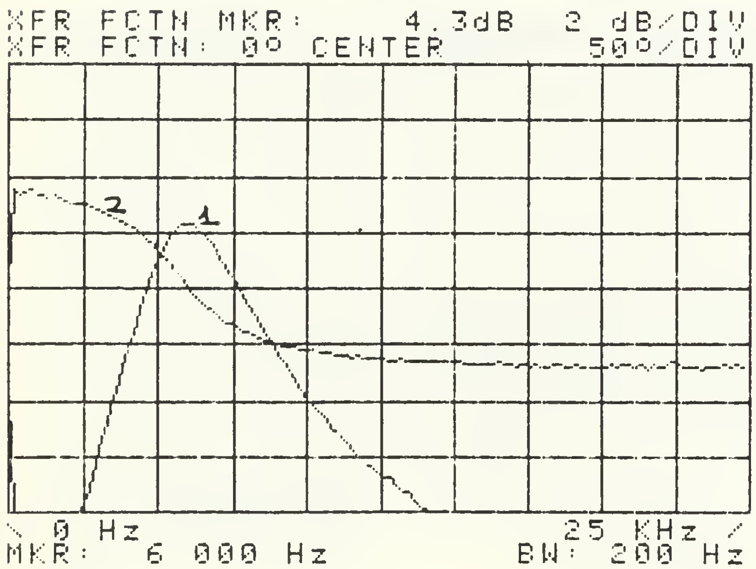
1 - Q = 10
2 - Q = 4
3 - Q = 1

Fig. 5.27 - BPF Amplitude Response
(R=1.6k), f=9.khz



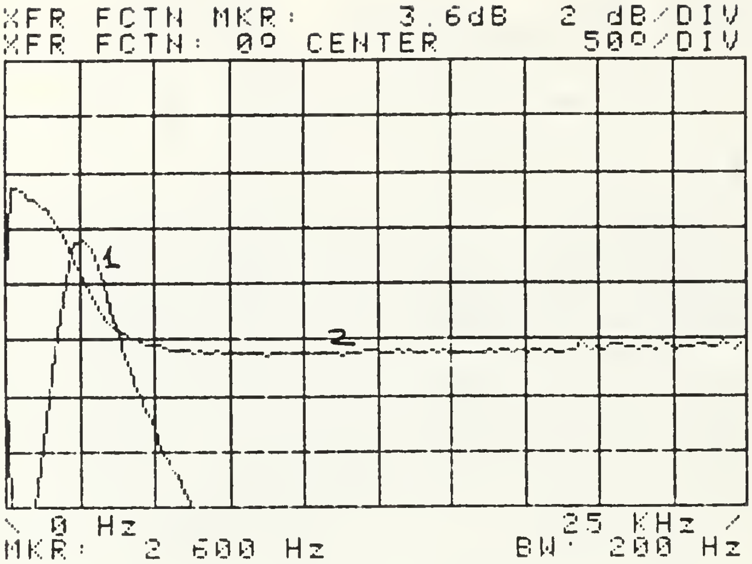
(a)

1. Phase Response
2. Amplitude Response



(b)

Fig. 5.8 - BPF Phase/Amplitude Response



1. Phase Response.
2. Amplitude Response.

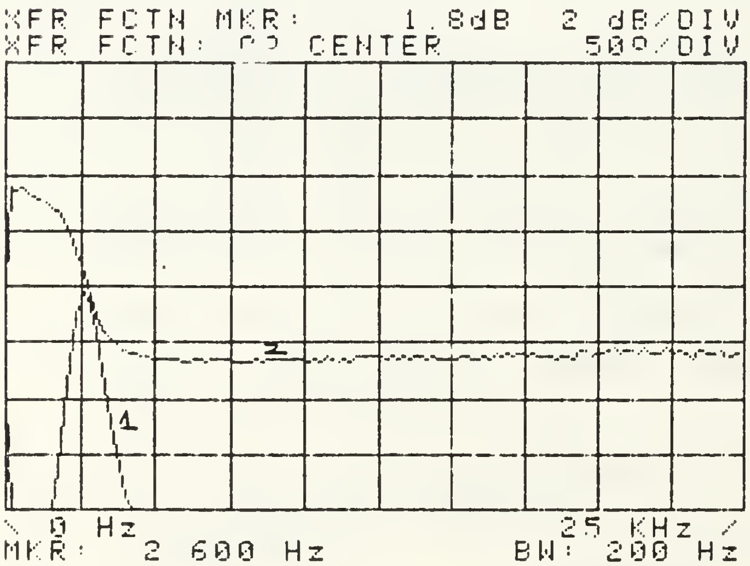
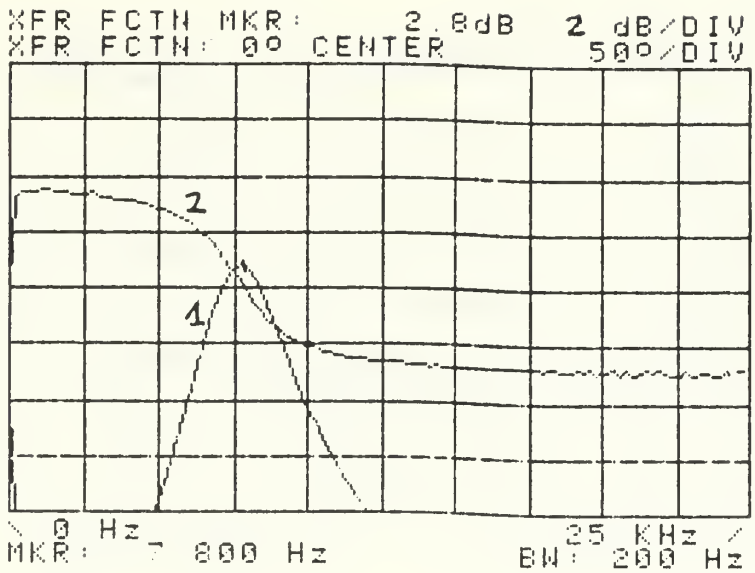
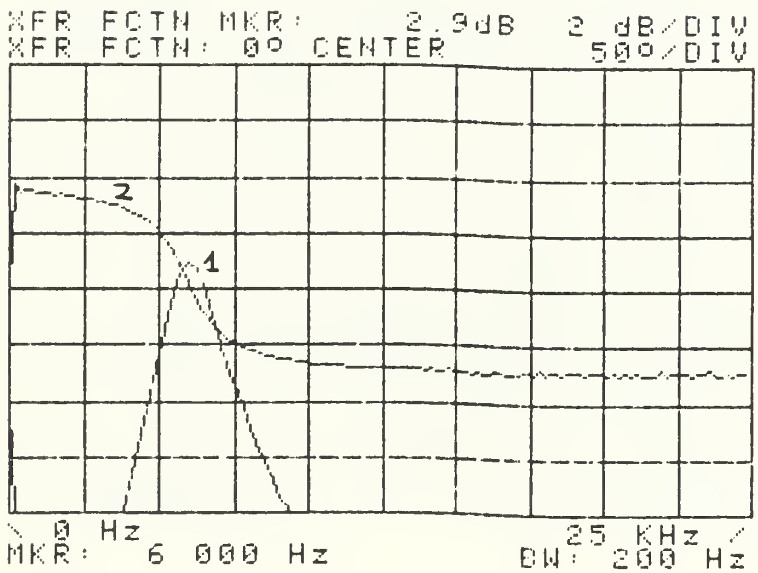


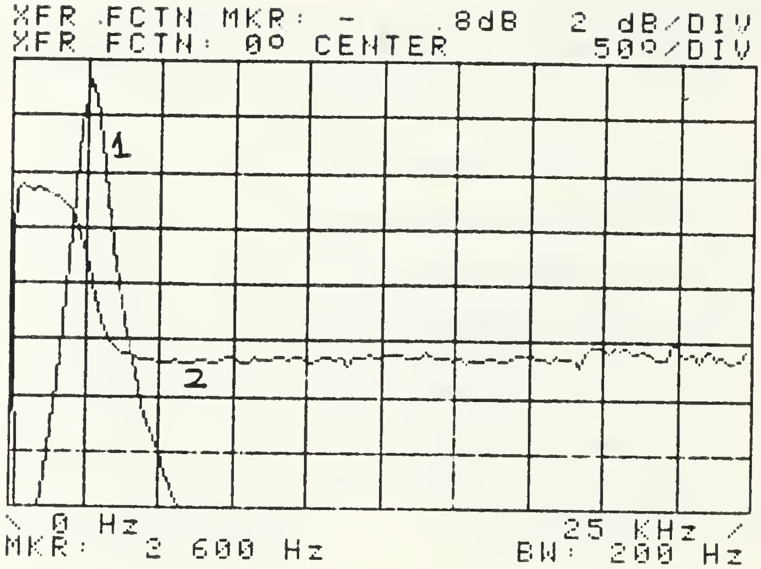
Fig. 5.29 - BPI Phase/Amplitude Response



1. Phase Response.
2. Amplitude Response.



5.30 - BPF Phase/Amplitude Response.



1. Phase Response.
2. Amplitude Response.

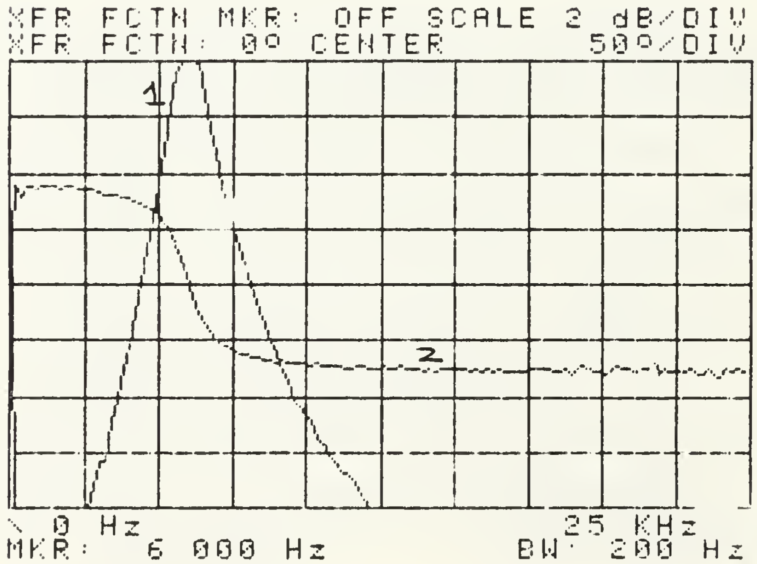
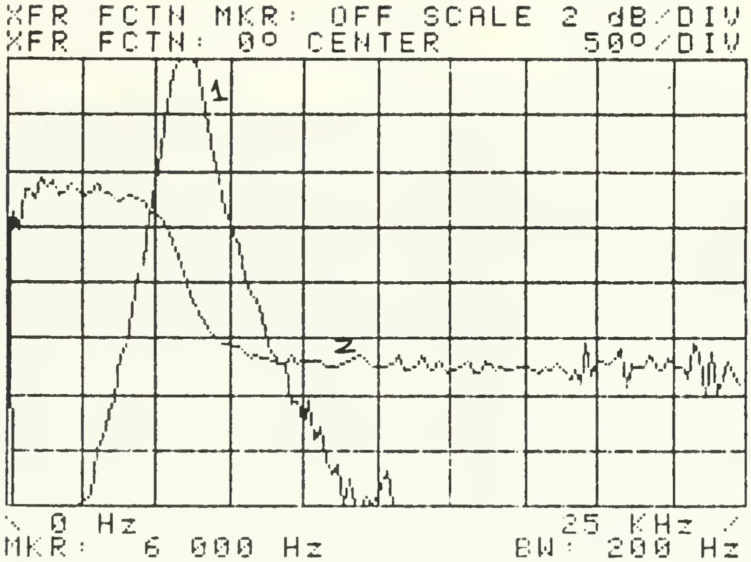


Fig. 5.31 - BPF Phase/Amplitude Response



1. Phase Response.
2. Amplitude Response.

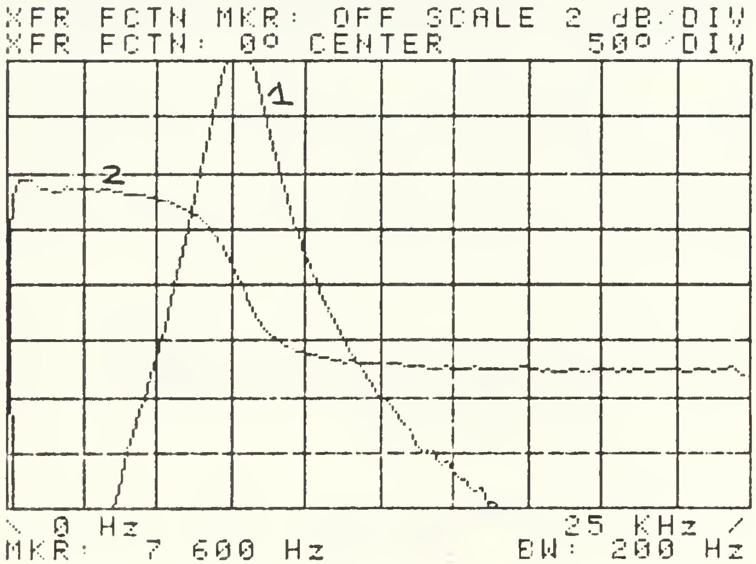


Fig. 5.32 - BPF Phase/Amplitude Response

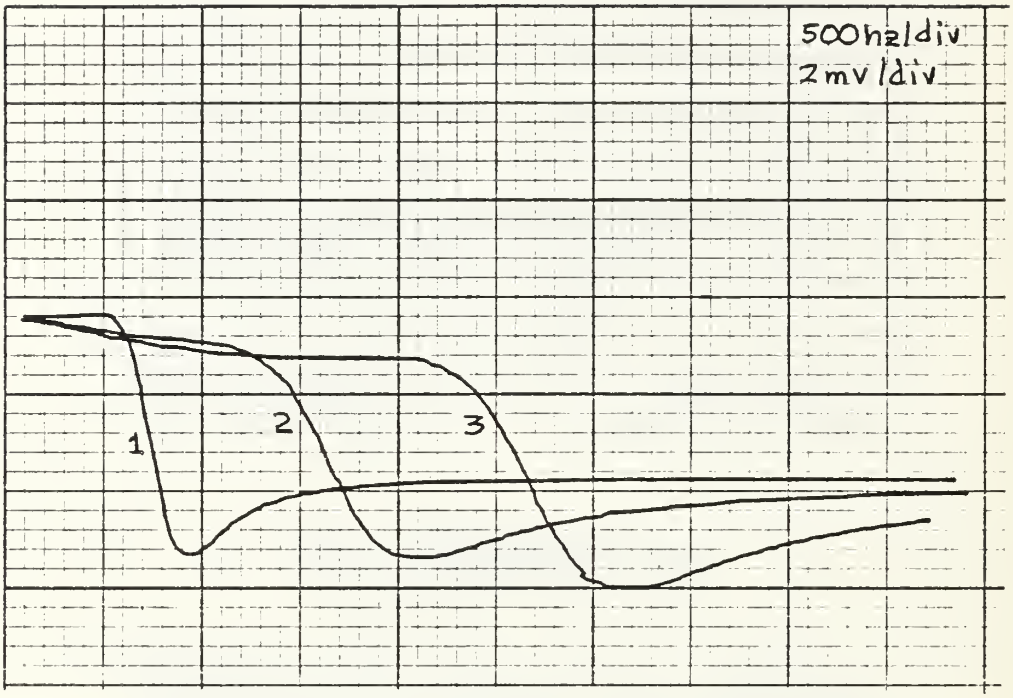
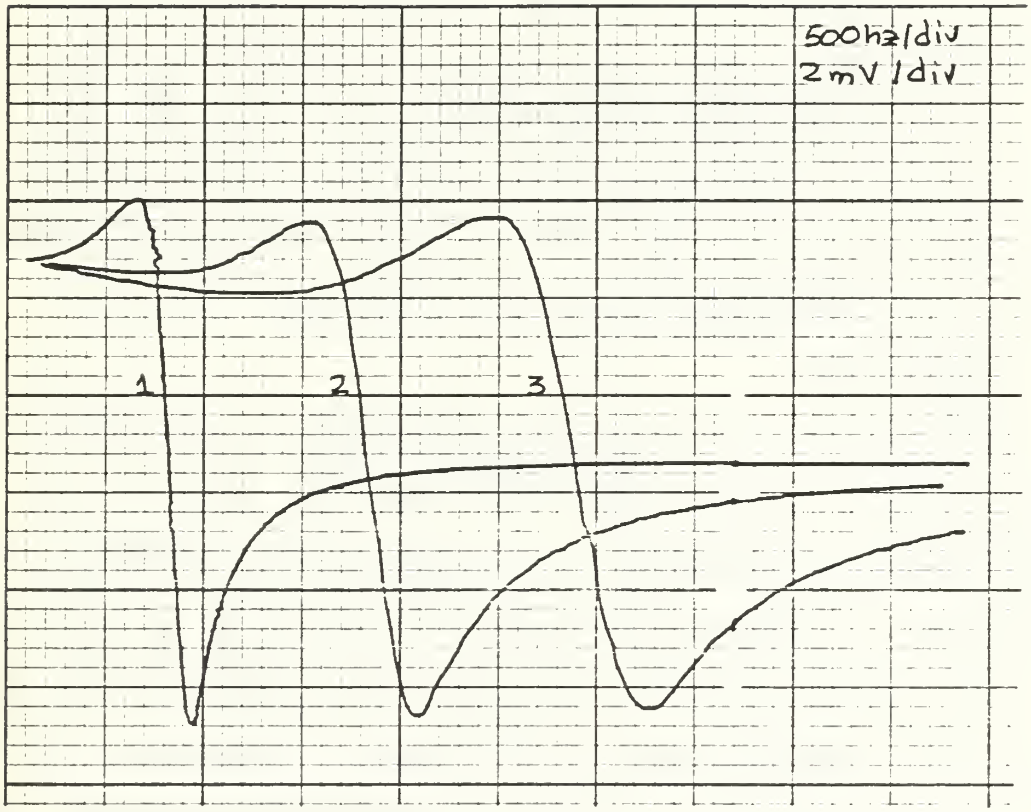


Fig. 5.33 - Notch Amplitude Response
(R=1.6k), Q=4



1 - f = 0.99khz
2 - f = 1.99khz
3 - f = 3.83khz

Fig. 5.34 - Notch Amplitude Response
(R=1.6k)

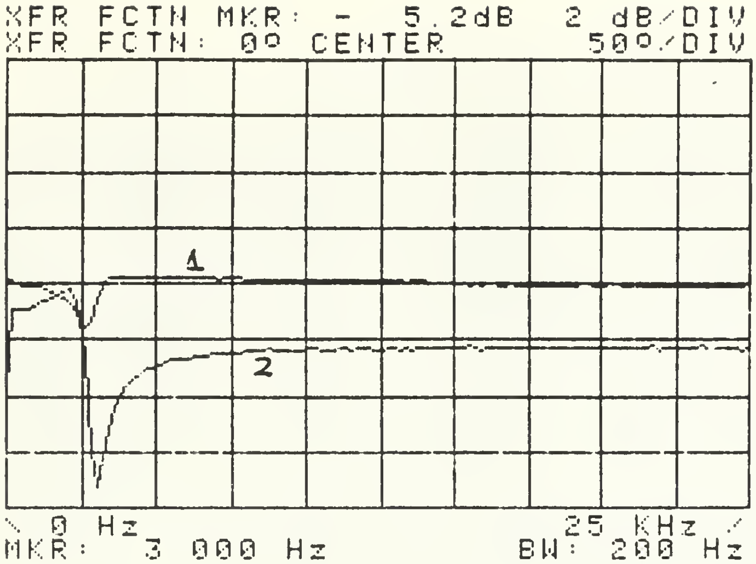
illustrate the amplitude response for different frequencies and for Q=4, while Figs. (5.35) and (5.36) illustrate the amplitude response for f=1KHZ and a variety of Qs. Fig. (5.37) illustrate the phase response in addition to the amplitude one.

5. <u>All Pass (AP) Realization</u>

The topology-control bitword 100 realizes the All-Pass filter. Figs (5.38), (5.39) and (5.40) illustrate the plotted amplitude and phase responses.

B. CONCLUSION

The constructed circuit performed as predicted by the theoretical analysis and the computer simulations. This means that it realized all the desired filtering transfer functions LP, HP, BP, N, and AP. The effect of interference of the control switches nonideal performances which is more severe at high Qps can be minimized by increasing the values of R (meaning that the values of Rq also increases and the values of Cs decreases).



1. Phase Response.
2. Amplitude Response.

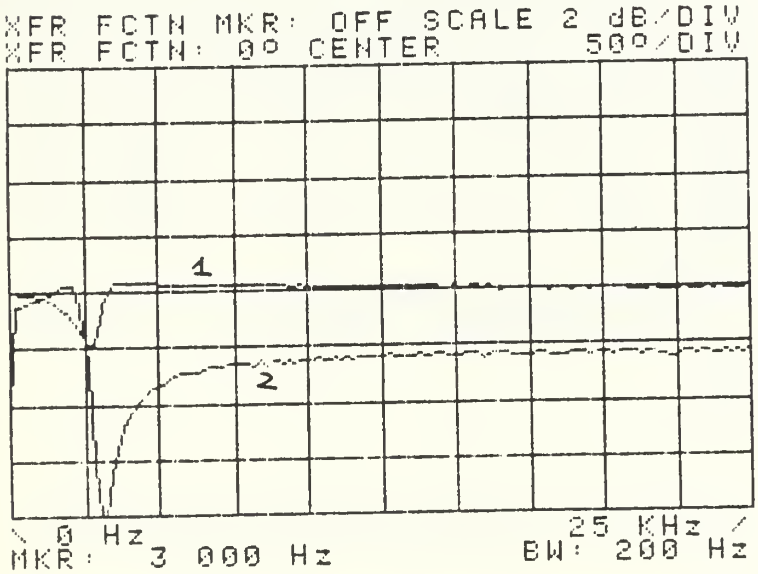
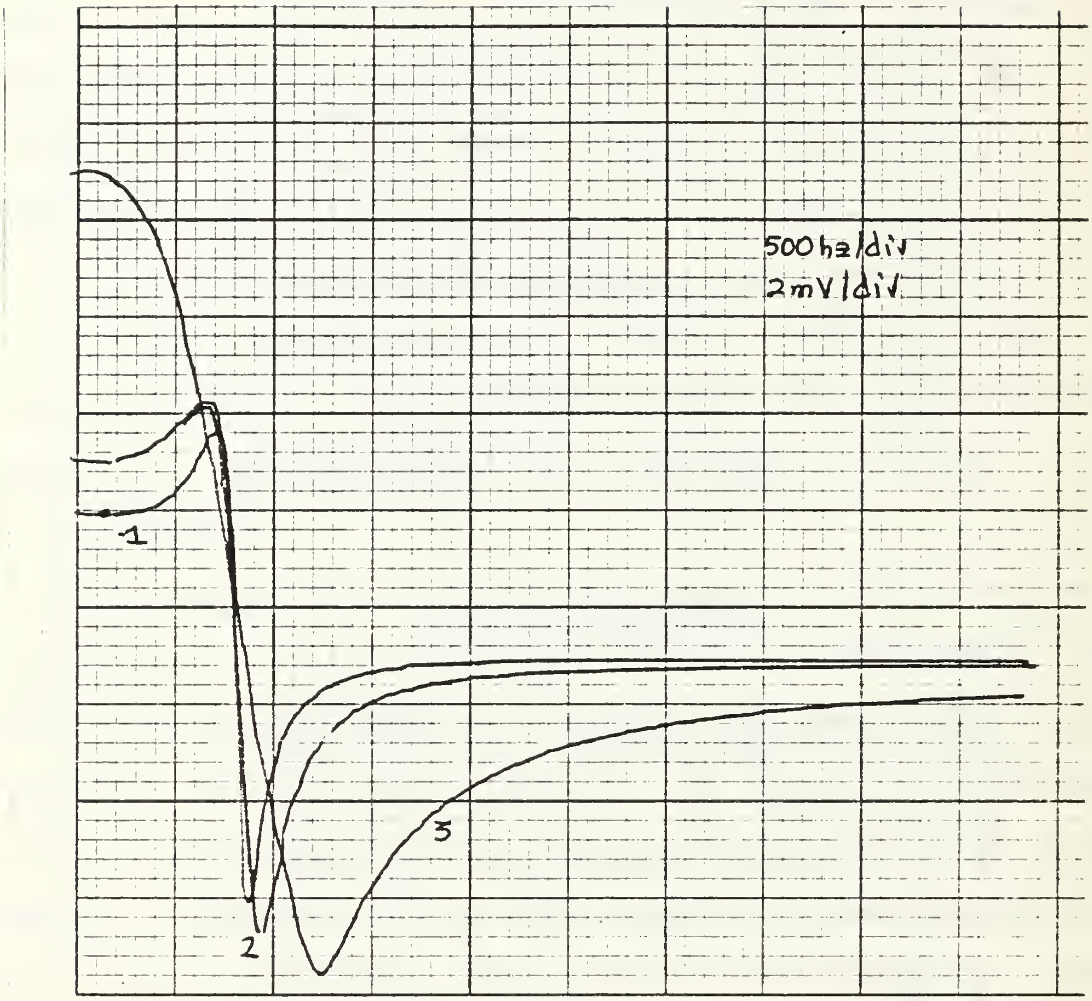
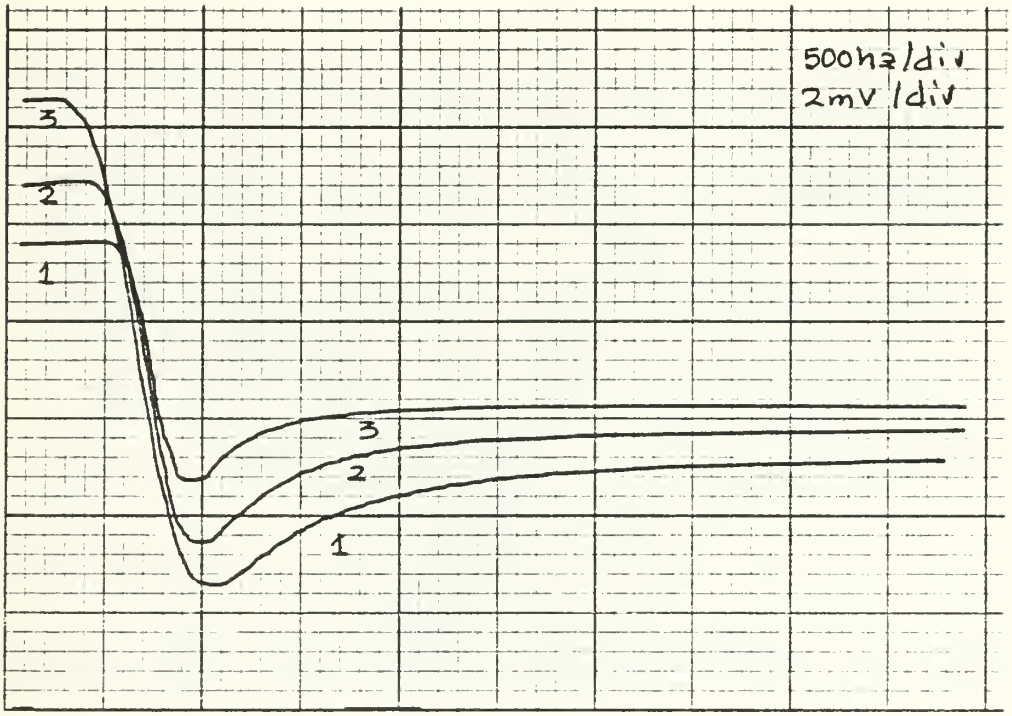


Fig. 5.35 - Notch Phase/Amplitude Response
f=2.8khz
(a) Q=7.5 (b) Q=5



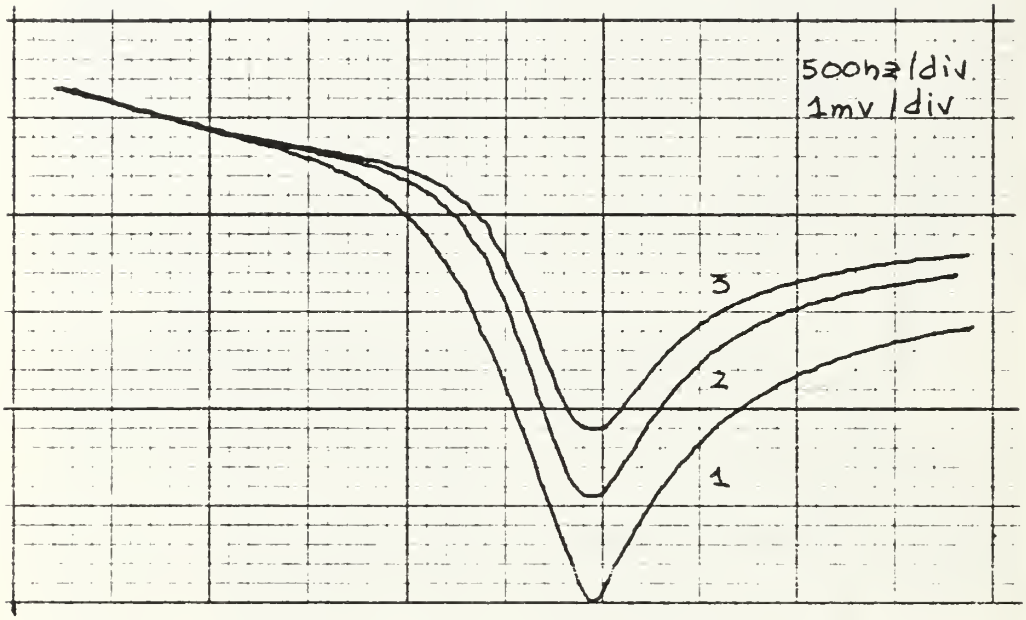
1 - Q = 2.0
2 - Q = 4.0
3 - Q = 8.0

Fig. 5.36 - Notch Amplitude Response
(R=1.6k), f=0.99khz



1 - Q = 2.0
2 - Q = 4.0
3 - Q = 8.0

Fig. 5.37 - Notch Amplitude Response
(R-1.6k), f=0.99khz



1 - Q = 4.0
2 - Q = 8.0
3 - Q = 12.0

Fig. 5.38 - APF Amplitude Response
(R=1.6k), f=2.8khz

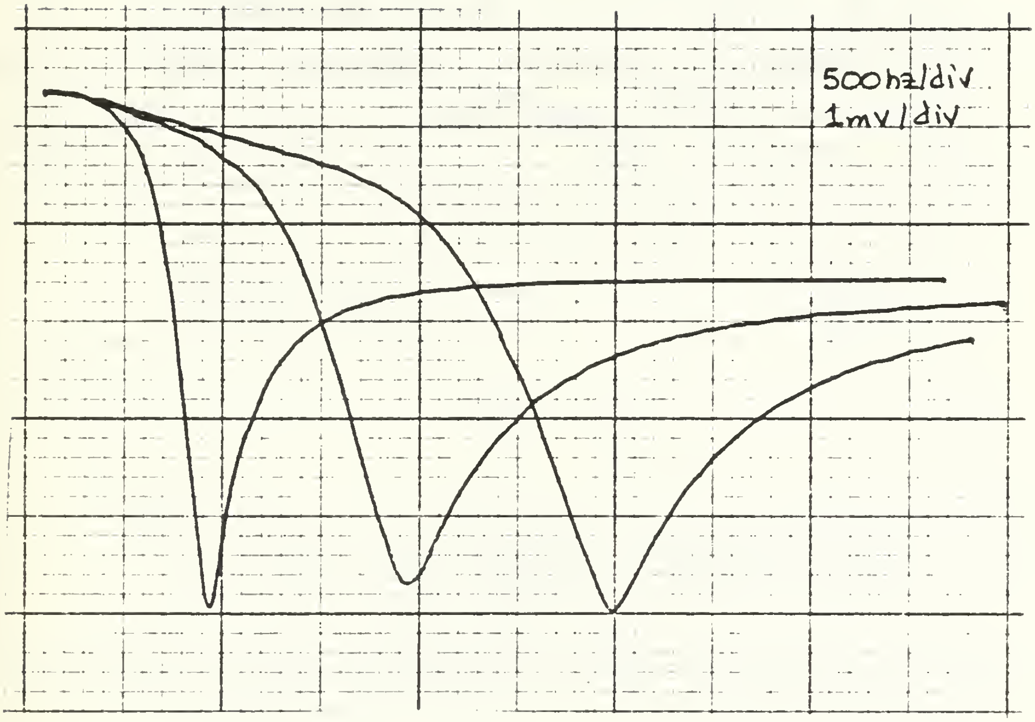
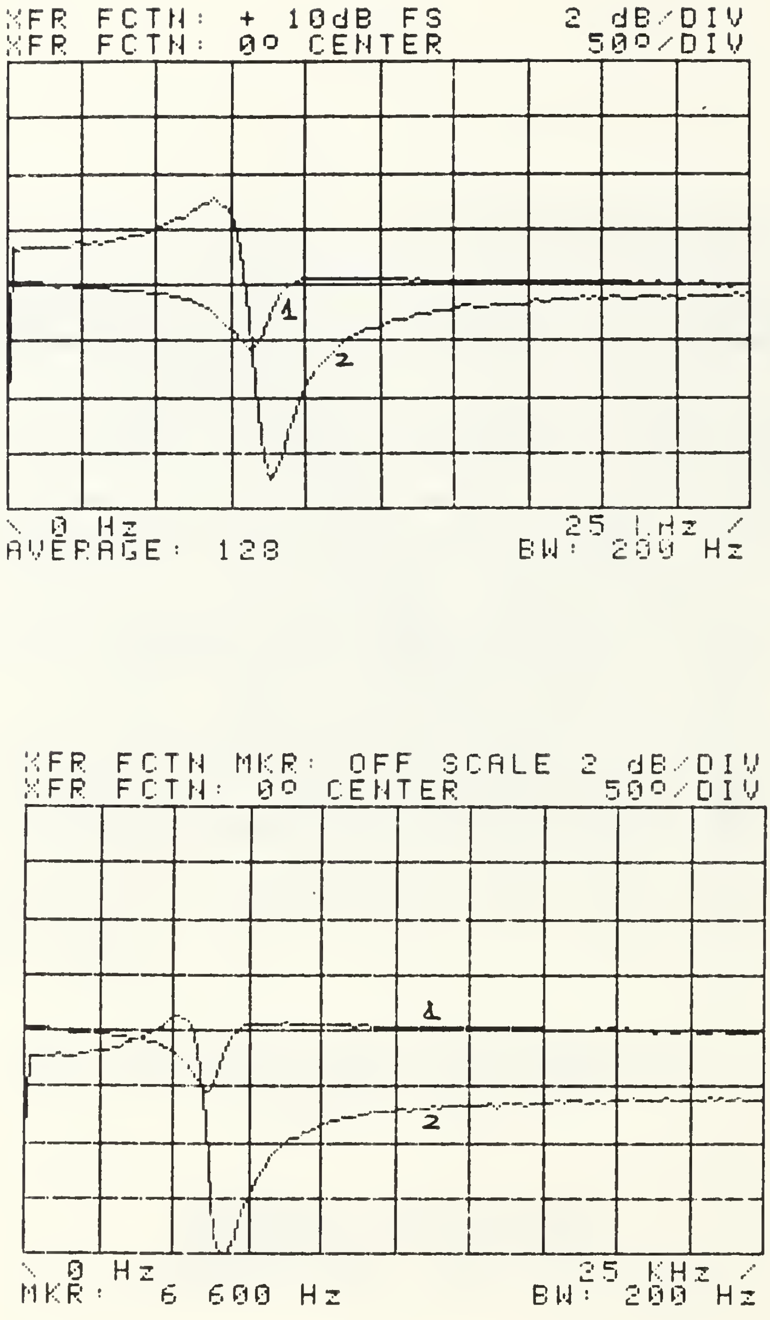


Fig. 5.39 - APF Amplitude Response
(R=1.6k) for f=0.99khz,
f=1.99khz, and f=2.8khz



1. Phase.
2. Amplitude

Fig. 5.40 - APF Phase/Amplitude Response (Q=7.5)

VI. COMBINING HIGHER ORDER SECTIONS

By cascading two or more programmable filters, higher order transfer functions can be obtained. Fig. (6.1) illustrates two cascaded GIC programmable filters.

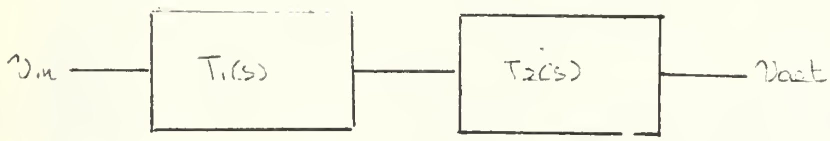


Fig. (6.1) Two Cascaded GIC Programmable Filters. (Each Black box Stands for the Network of Fig. (3.10)

25 different combinations of the two individual transfer functions Ti(s) and Tj(s) can obtained as it is indicated in Table (4.3).

For Lp-LP combination and for ideal theoretical case (A1=A2-->oo) a fourth order low pass filter can be obtained with transfer function:

$$T_2(s) = \frac{\omega_p^4}{s^2 + \frac{\omega_p}{Q_p} s + \omega_p^2} \qquad (6.1)$$

The computer simulation of the fourth order transfer function of a nonideal theoretical filter vs. the second order case is illustrated at Fig. (6.2). The experimental results taken from the properly designed and built circuit as illustrated in Fig. (6.3) are shown in Fig. (6.4).

Using the same procedure as above a fourth order HP-HP combination for the ideal theoretical response is given by

$$T_1(s) = 4\frac{s^4}{\left\{\omega p^2 + \frac{\omega p}{Q_p} s + s^2\right\}^2} \tag{6.2}$$

which at Wp takes the complex value of

$$T_1(j\omega_p) = -4jQ_p^2 \tag{6.3}$$

with magnitude of

$$|T_1(j\omega_p)| = 4Q_p^2 = 40\log(2Q_p), db \tag{6.4}$$

Fig. (6.5) illustrates the fourth order nonideal theoretical HP filter response vs. the second order one. Both at Qp=2 and fp=8KHZ, while fig. (6.6) illustrates the experimental responses. For Qp=2 according to (6.4) and

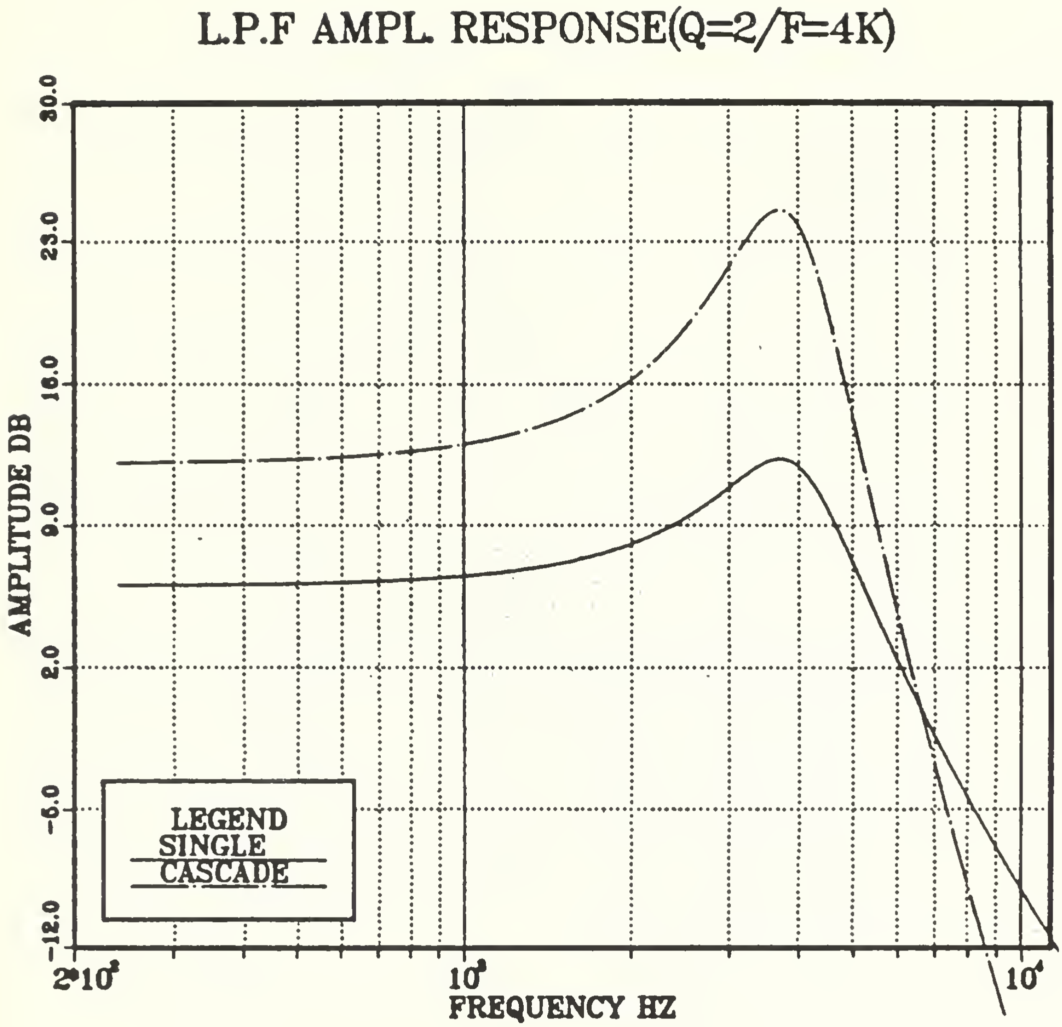
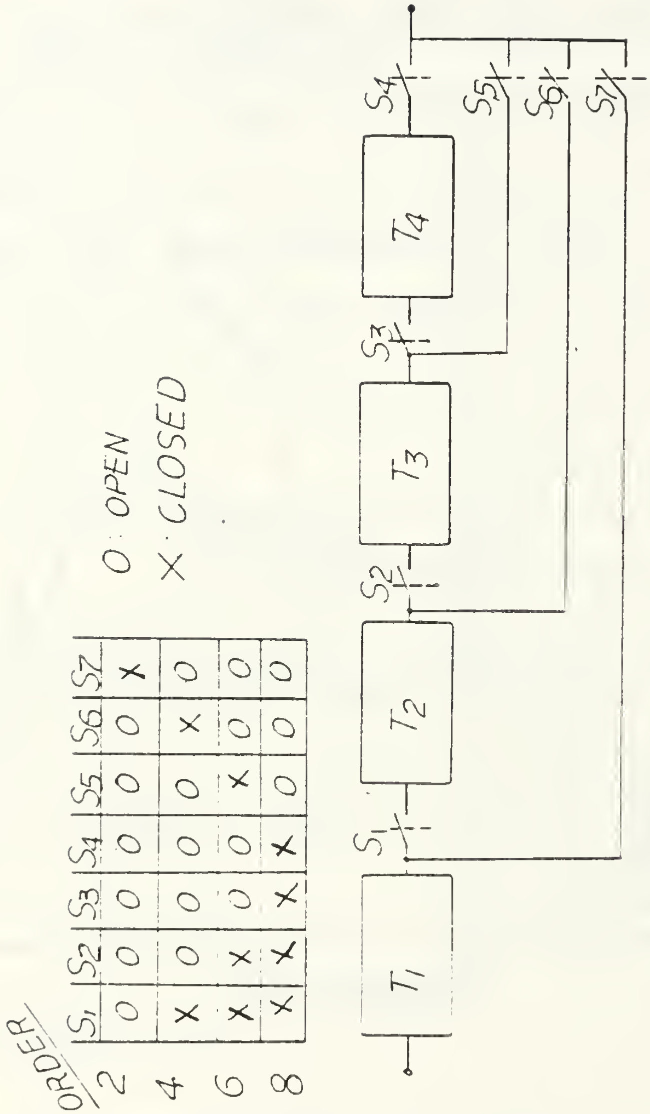
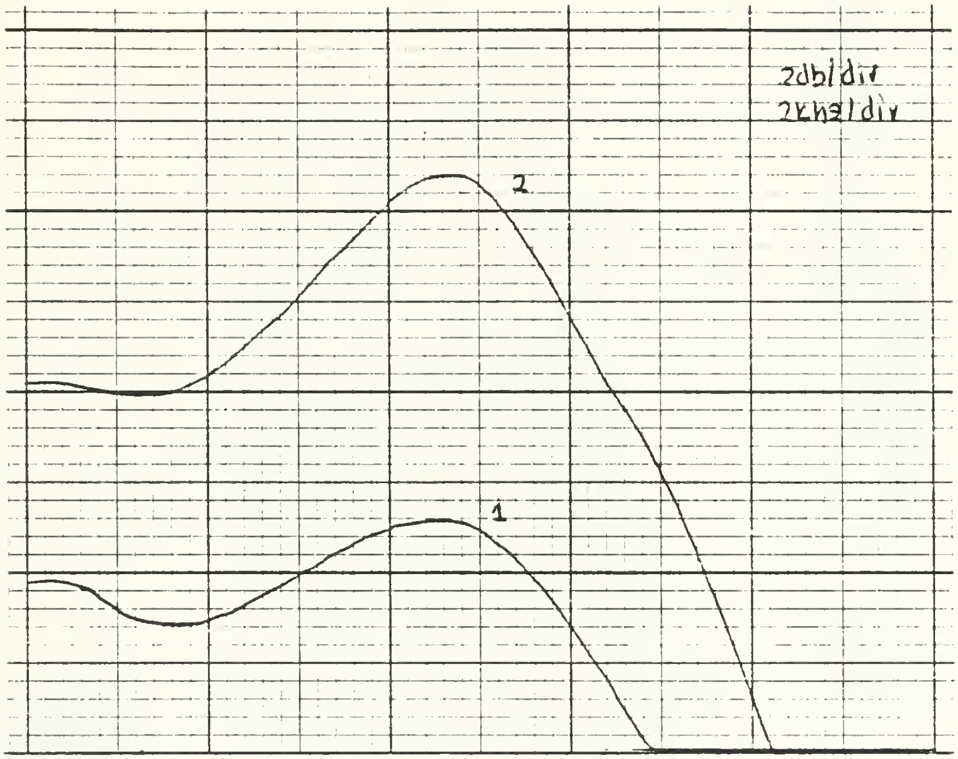


Fig. 6.2 - LPF Fourth Order vs. Second Order
Ideal Response from Computer Simulation.



ORDER	S1	S2	S3	S4	S5	S6	S7
2	O	O	O	O	O	O	X
4	X	O	O	O	O	X	O
6	X	X	O	O	X	O	O
8	X	X	X	X	O	O	O

Fig. 6.3 - Network with Logic for Realizing np to 8th Order LP, HP, BP, N, and AP Transfer Functions.



1. Single
2. Cascade

Fig. 6.4 - 4th Order vs. 2nd Order Experimental LPF Response (Q=2)

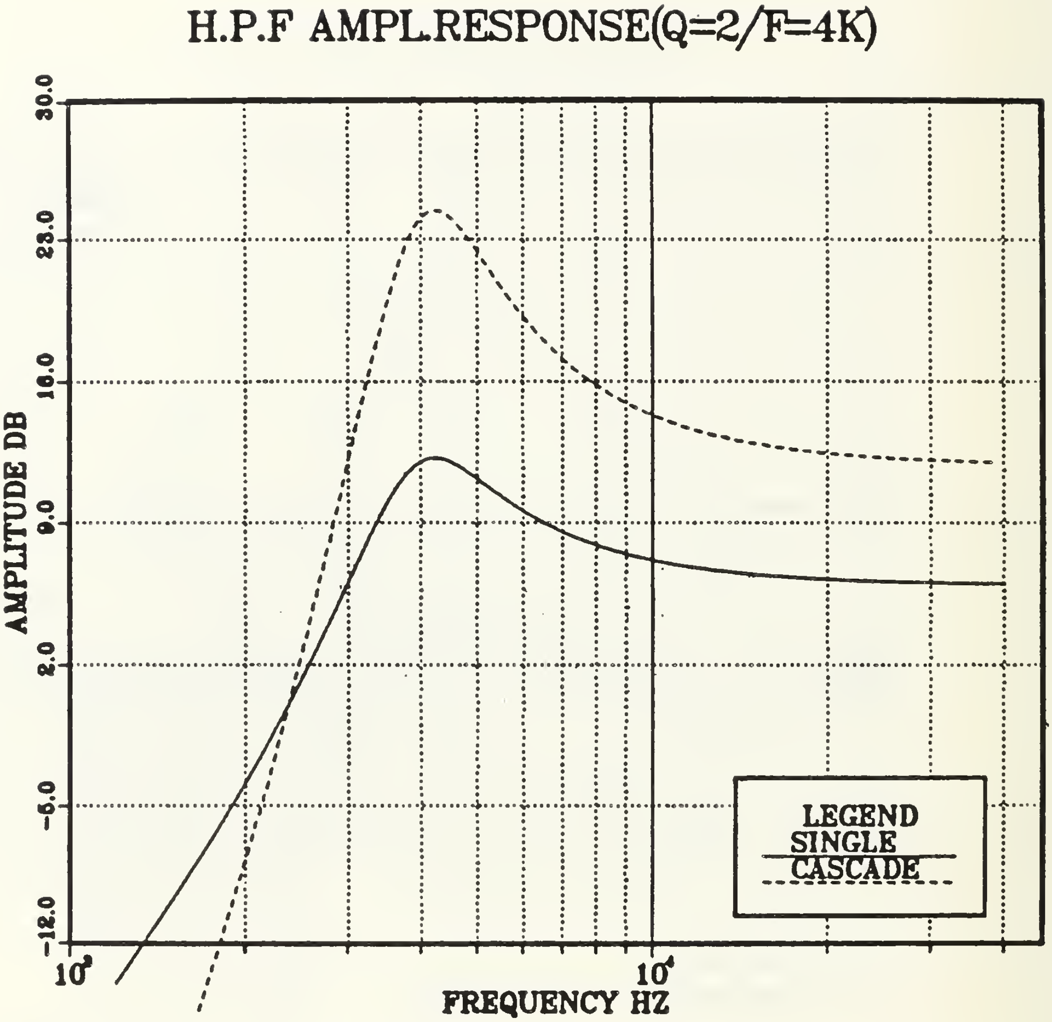


Fig. 6.5 - "Ideal" 4th vs. 2nd Order HPS Responses From Computer Simulation.

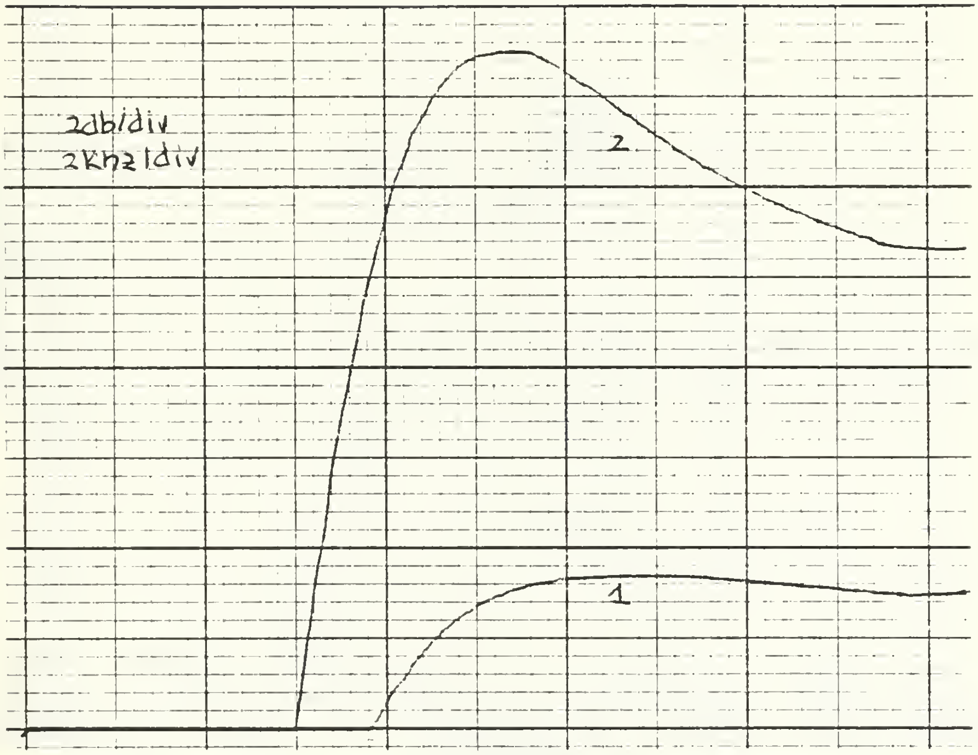


Fig. 6.6 - 4th Order vs. 2nd Order HPF obtained Responses from the Constructed Circuit.

(2.4) a difference of 12dB were expected between the second and fourth order filter which is approximately the case in both computer simulation and experimental results.

The BP-Bp combination leads to the theoretical "ideal" fourth order transfer function:

$$T_1(s) = 4 \frac{\left(\frac{\omega_p}{Q_p}\right)^2 s^2}{\left\{s^2 + \frac{\omega_p}{Q_p} s + \omega_p^2\right\}^2} \tag{6.5}$$

which takes the following value at ωp:

$$T_1(j\omega_p) = 4 = 12db \tag{6.6}$$

According to this, a magnitude response of 12 dB approximately at ωp, was expected from both experimental and computer simulation results. Fig. (6.7) illustrates the simulated response of a fourth order nonideal HP filter vs. a second-order one. A difference of 5.8 dB instead of 6 dB can be observed. Fig. (6.8) illustrates the experimental response, where a difference of 5.4 dB appears basically due to the interference of the control switches and the nonideally matched values of the capacitors which control the fp selection.

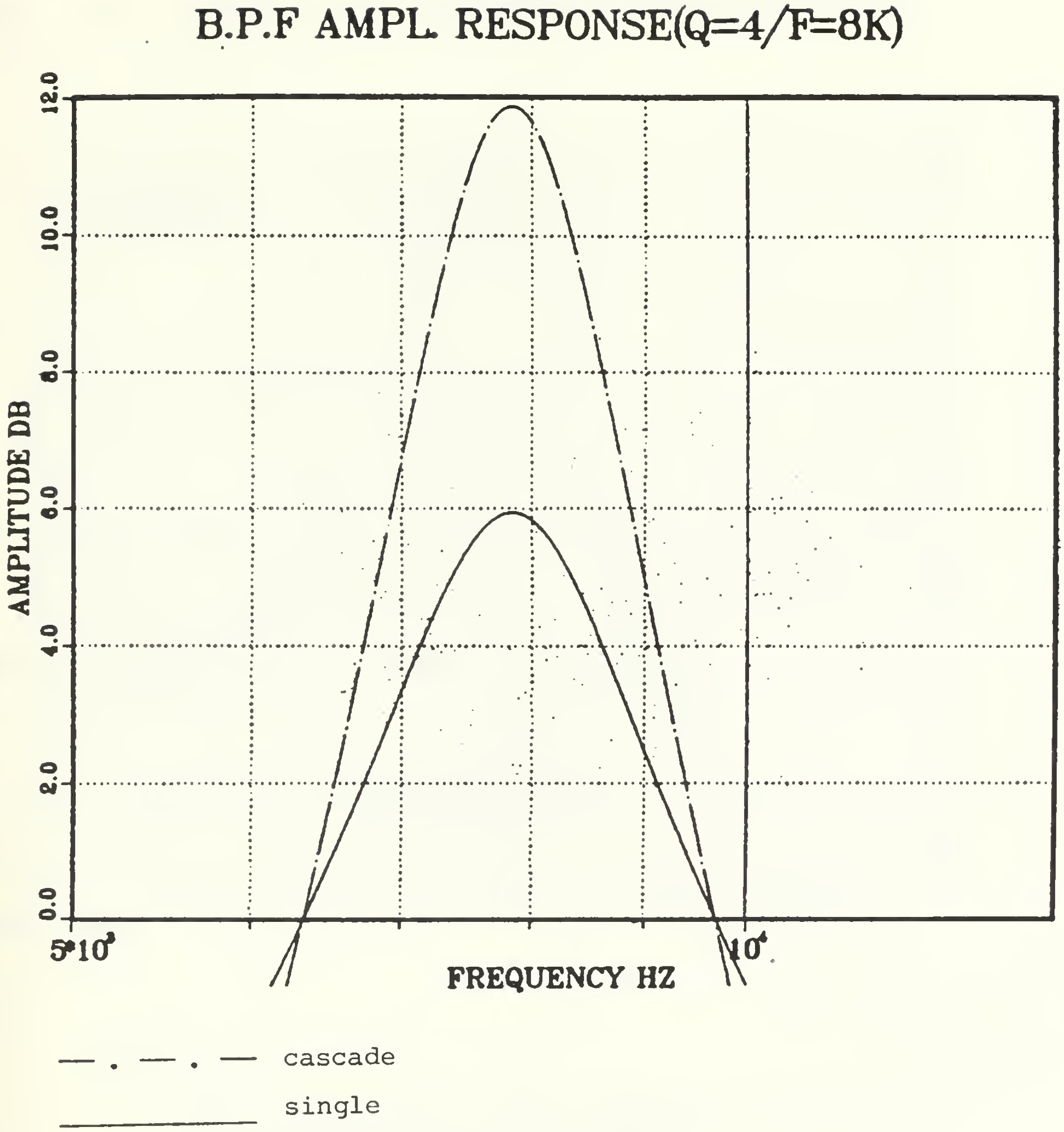


Fig. 6.7 - Fourth Order vs. Second Order BPF Amplitude Responses from Computer Simulation.

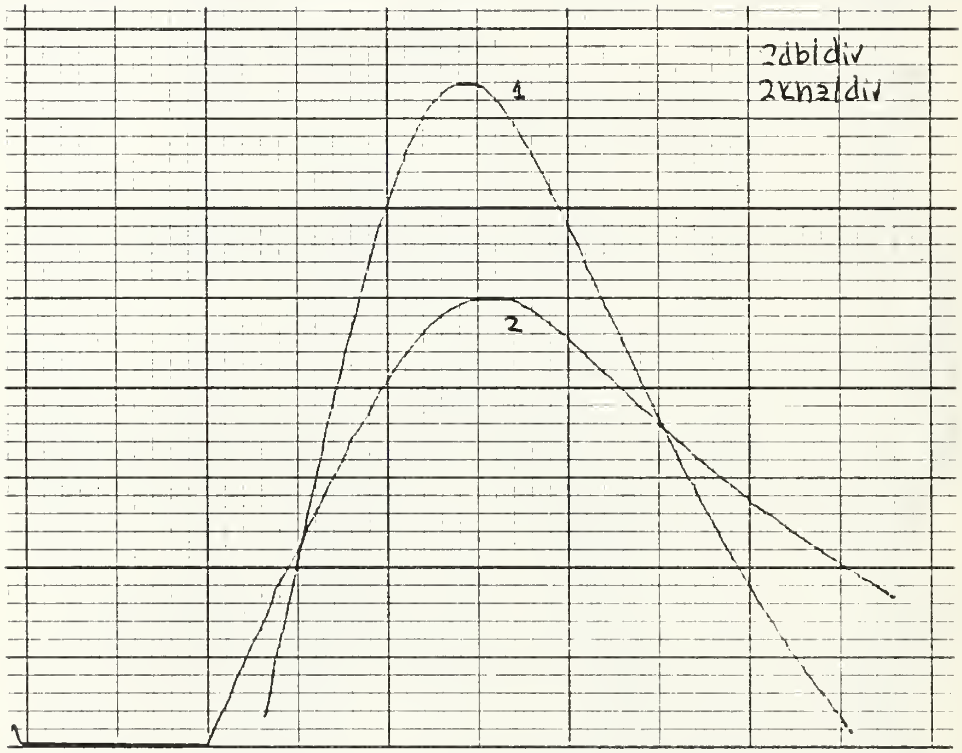


Fig. 6.8 - Fourth vs. Second Order BPF
Responses Obtained from the
Constructed Circuit.

The N-N combination for the ideal theoretical case (A1=A2-->oo) results the following transfer function:

$$T_2(s) = \frac{(s^2+\omega_n)^2}{\left\{s^2+\frac{\omega_p}{Q_p}s+\omega_p\right\}^2} \tag{6.7}$$

where ωn is the Notch frequency.

Fig. (6.9) illustrates the fourth order "nonideal" Notch filter simulated response vs. the second order one.

The AP-Ap combination for the theoretical ideal case result in the following transfer function:

$$T_1(s) = \frac{\left\{s^2-\frac{\omega_p}{Q_p}s+\omega_p^2\right\}^2}{\left\{s^2+\frac{\omega_p}{Q_p}s+\omega_p^2\right\}^2} \tag{6.8}$$

which takes the following values for S=0 and oo:

$$T_1(j0) = 1 \tag{6.9}$$

with amplitude and phase of

$$|T_1(j0)| = 1 = 0db, \quad \underline{/T_1(j0)} = 360^\circ \tag{6.10}$$

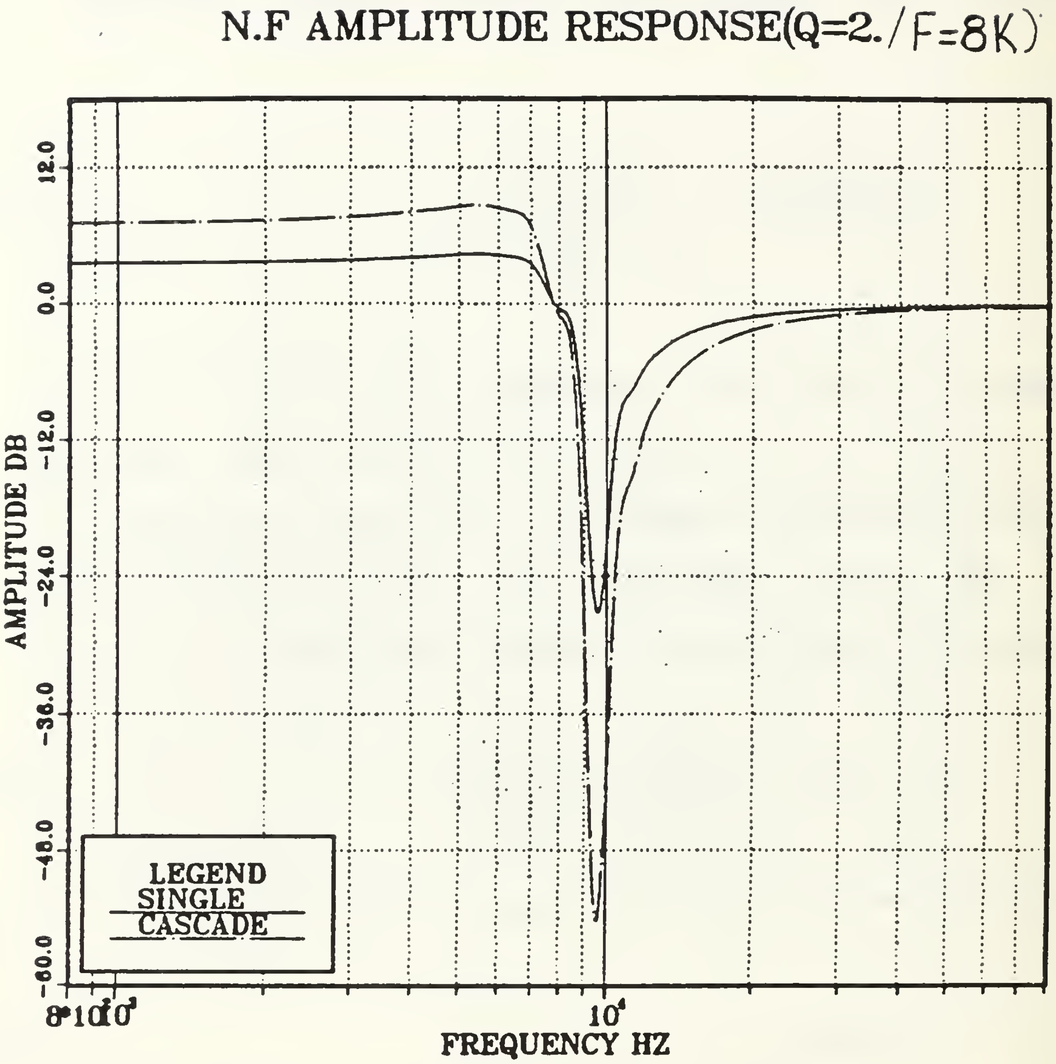


Fig. 6.9 - Fourth Order vs. Second Order Notch Filter Amplitude Response from Computer Simulation.

and

$$T_1(j\infty) = 2.25 \tag{6.11}$$

with amplitude and phase of

$$|T_1(j\infty)| = 7db \ , \ \underline{\angle \ T_1(j\infty)} = 0^{\circ} db \tag{6.12}$$

The computer simulation of nonideal fourth order all pass filter vs. a second order one which is illustrating at Fig. (6.10) matches the above.

Fig. (6.11) illustrates the resulting Chebychev filter from a BP-BP combination with Qp=4 and different frequencies (fp1=8KHZ fp2=10KHZ), while Figs. (6.12) and (6.13) illustrate the resulting Chebychev filters from the designed and built circuit. Fig. (6.14) illustrates the resulting response from a HP-LP combination.

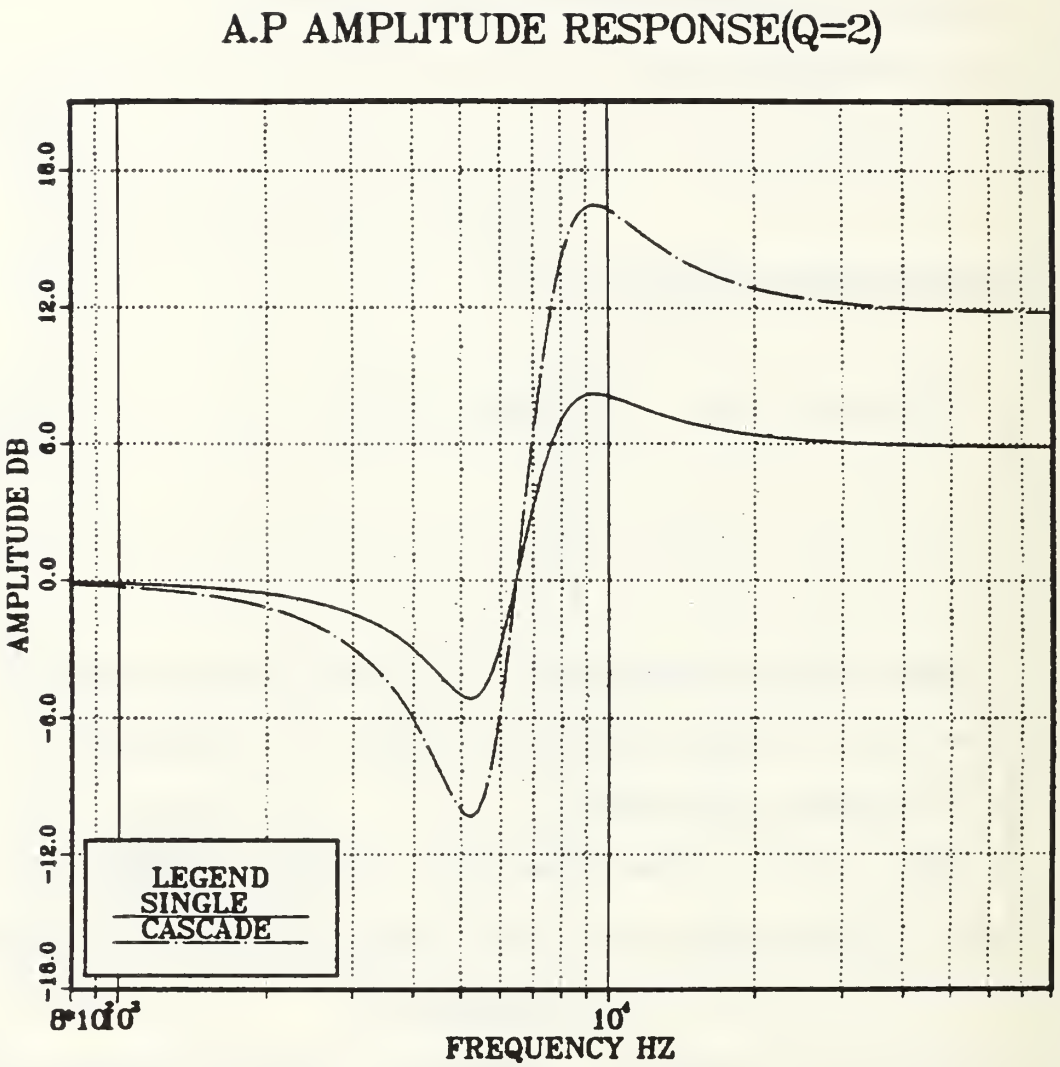


Fig. 6.10 - APF Fourth Order vs. Second Order Amplitude Response from Computer Simulation.

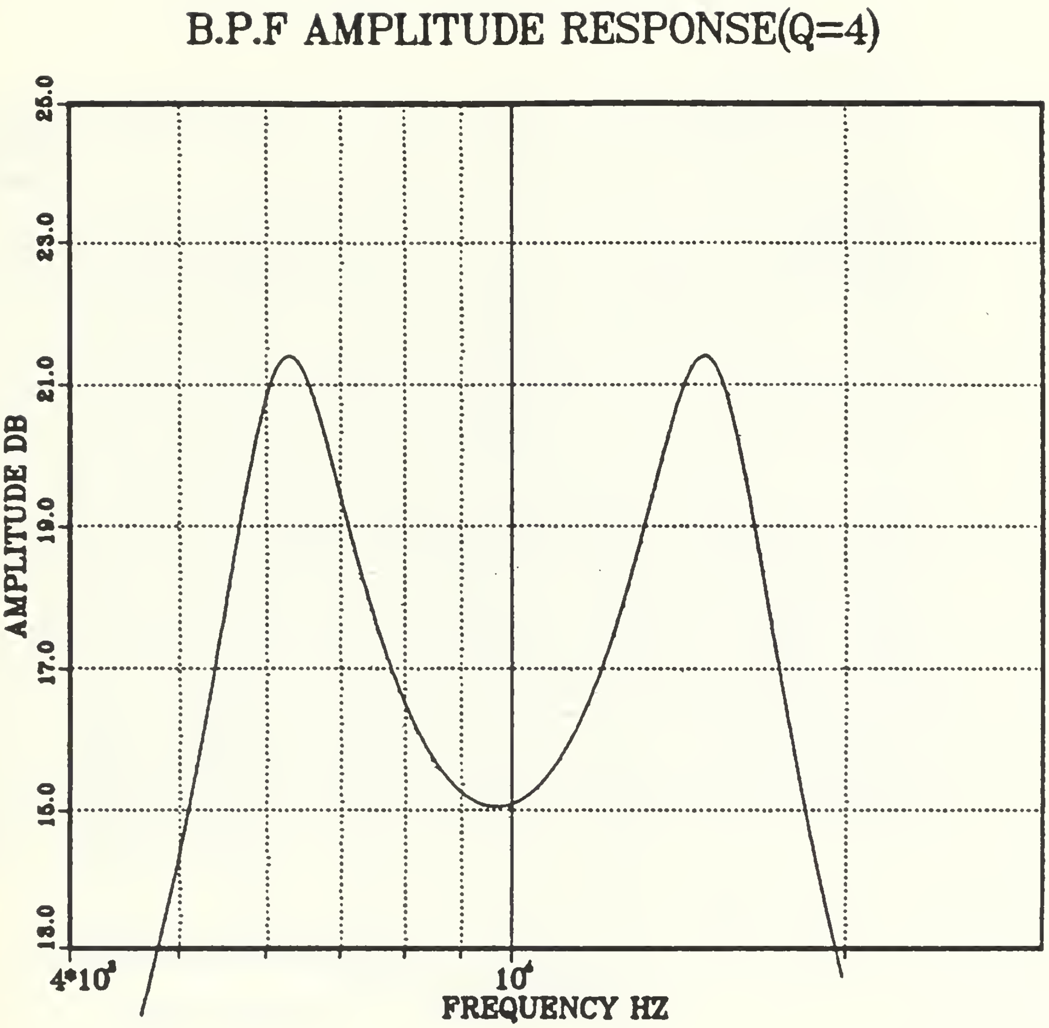


Fig. 6.11 - Chevycheb Response for f_1=8k,
f_2=10k

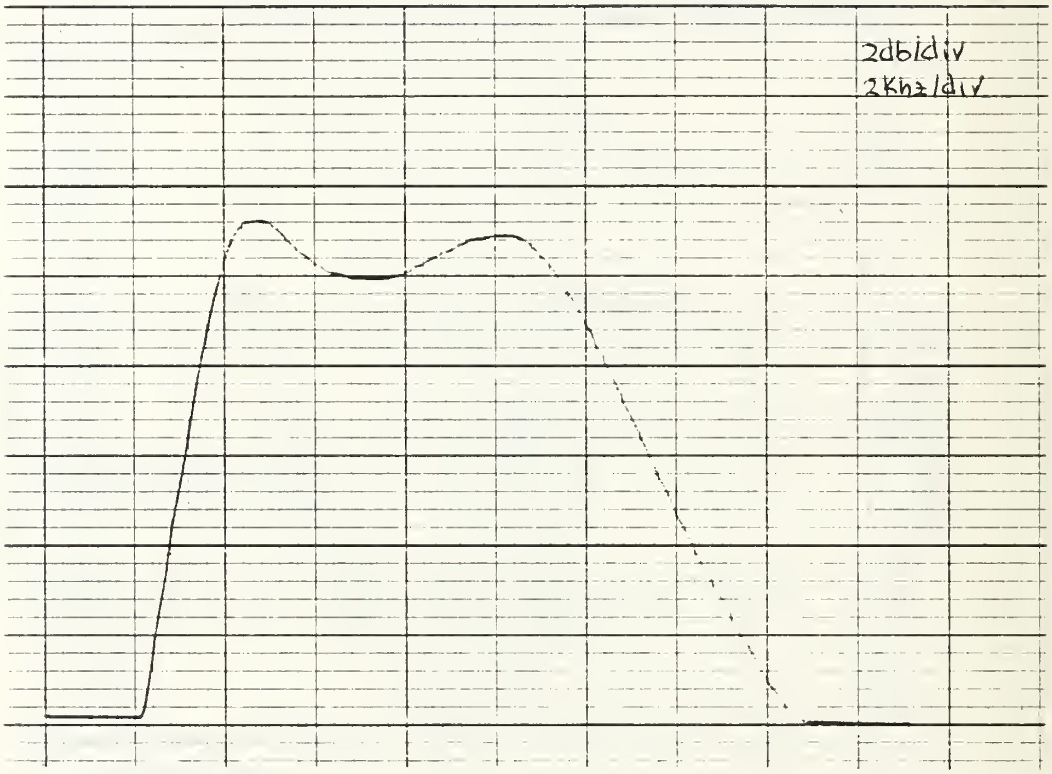


Fig. 6.12 - Chevychev Response Obtained by Cascading BP-BP
f_1=6.65khz f_2=12.8khz

Q_1=3 Q_2=3.5

Fig. 6.13 - BP-BP Response $Q_1=Q_2=4$

(f_1=3khz, f_2=15khz)

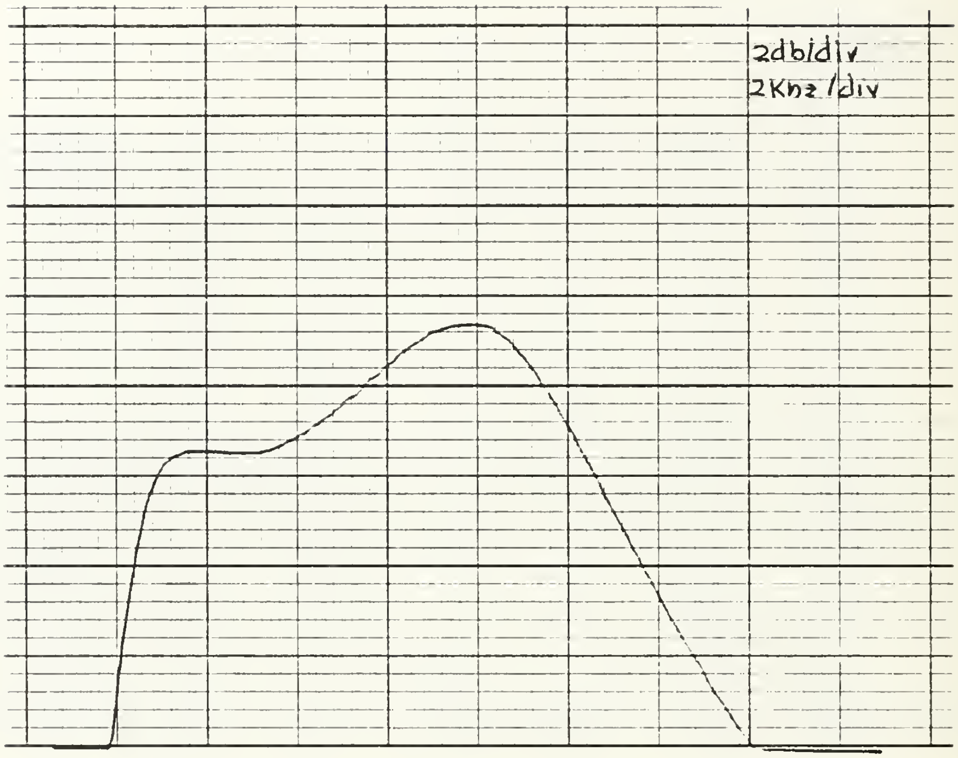


Fig. 6.14 - HP-LP Response $Q_1=Q_2=4$

(f_1=3khz, f_2=9.0khz)

VII. APPLICATION OF THE PROPOSED GIC PROGRAMMABLE FILTER IN FREQUENCY HOPPING SYSTEMS

A. BACKGROUND

1. General Description of Frequency Hopping Signals

Frequency hopping is a spread spectrum modulation technique used to generate many possible carrier frequencies over a large bandwidth. Of all the possible carrier frequencies, only one is selected at a given time. However, all frequencies are eventually selected during some time interval.

Frequency hopping (FH) may be pictured as an RF carrier whose center frequency is "hopped" over many frequencies. The hopping may be either in a simple sequence or a pseudorandom sequence.

The hopping rate of a frequency hopping system does not affect the bandwidth of the output spectrum. In a direct sequence system the chip rate determines the total bandwidth. In a frequency hopping system, however, the bandwidth is determined by the highest and lowest frequencies of the frequency hopped carriers. For example, if the highest frequency carrier is at 15 MHz and the lowest frequency carrier is at 10 MHz, the total signal bandwidth is 5 MHz. This is the bandwidth regardless of the hopping rate. This allows wideband spread spectrum signal generation at low hopping rates.

2. <u>Signal Generation</u>

Frequency-hopped signals may be generated in several ways. The different methods are classified into two groups:

(1) Direct synthesis, and

(2) Indirect synthesis.

One important aspect of frequency hopping synthesis is coherency. coherent signal synthesis is defined as the establishment of a known and repeatable phase each time a new frequency is hopped to. Non-coherent signal synthesis is defined as the establishment of a random or unknown phase each time a new frequency is hopped to. some techniques, direct or indirect, of signal generation can be used as a coherent frequency source. In other techniques, the changing of frequencies creates non-coherent sources.

If a frequency hop system is a coherent, it will have a signal-to-noise advantage over a non-coherent system.

a. Direct Synthesis

The direct approach to signal synthesis utilizes techniques which enable direct synthesis of different frequencies. Examples of direct synthesis techniques are:

(1) Frequency mixing, and

(2) Surface acoustic wave devices.

Frequency mixing for single synthesis is a common technique used to generate many different frequencies. An example of the frequency mixing technique is show in Fig. (7.2).

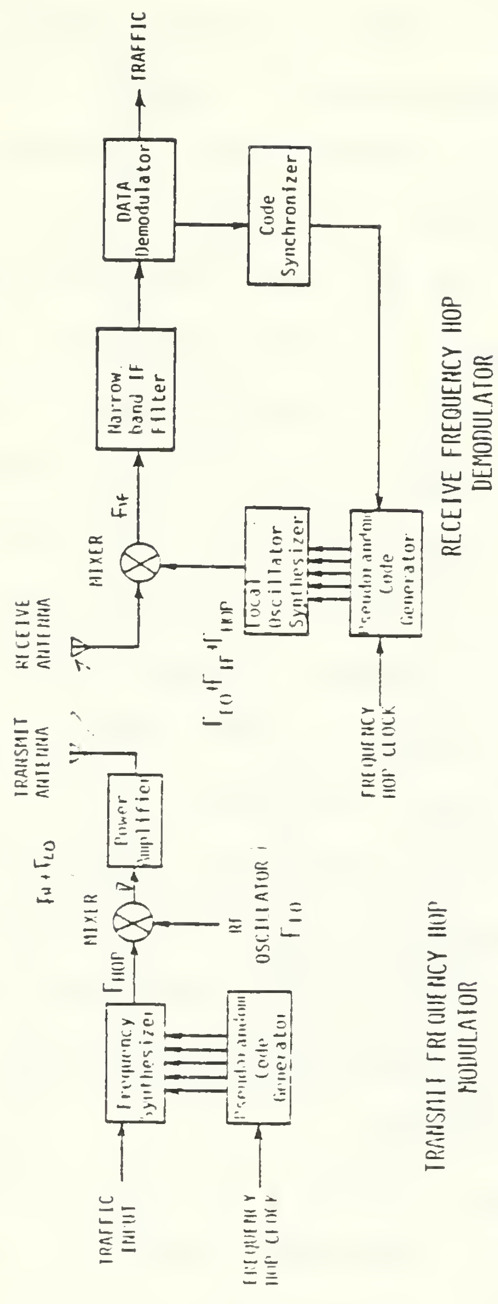


FIGURE 7.1 BLOCK DIAGRAM OF A FREQUENCY HOP MODEM

A pseudorandom code generator selects one of many transmit frequencies during a small time interval. The traffic modulates the carrier frequency which is spread over many different frequencies by the hopping action. The receiver dehops the input signal into a narrowband IF. The code synchronizer locks the pseudorandom code generator in the receiver to the received signal. A data demodulator removes the traffic from the IF amplifier output.

In Fig. (7.2), an RF switch selects one of several frequency inputs. Two of these input signals of different frequencies are multiplied together to generate a new output frequency. The device used to multiply the two signals together is called a frequency mixer. when two frequencies are mixed, the sum and the difference of the frequencies are generated. In order to select only one of these frequencies, a "filter is used to reject the unwanted frequency." A filter tuned to the desired frequency would allow selection of that frequency while rejecting the other. By selecting the mixing frequencies in the proper order, the output frequency can be stepped through several different frequencies. At each frequency mixer output, a filters is required to reject unwanted frequencies. The filters may require a short time for the signal to stabilize after it is selected. The time required for the filter to stabilize at each new frequency may ultimately determine the maximum hopping rate of the direct frequency synthesizer.

b. Indirect Synthesis

The indirect method of signal synthesis is defined as frequency synthesis through the use of phase-locked oscillators. One common indirect method for synthesis is shown in Fig. (7.3).

In this circuit a phase-locked loop is used to generate the numerous carrier frequencies. The phase-locked loop has an internal oscillator whose output frequency, Fo,

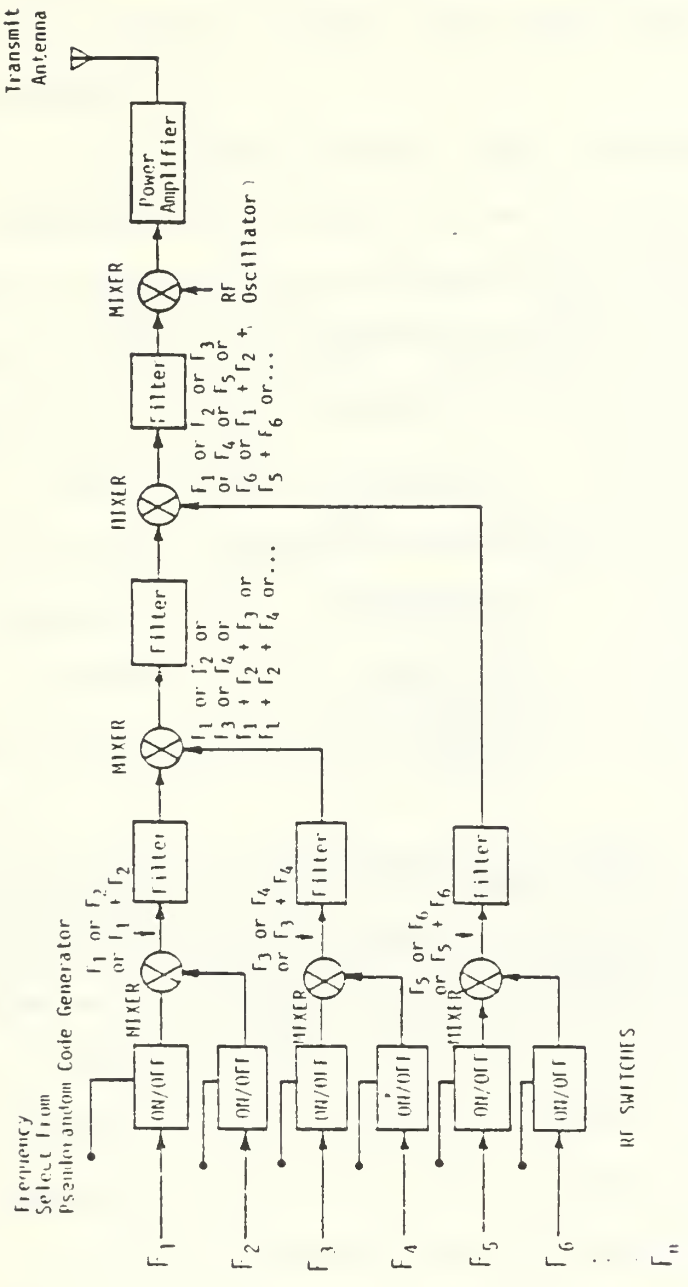


FIGURE 7.2 DIAGRAM OF A DIRECT FREQUENCY SYNTHESIZER

A pseudorandum sequence selects different combinations of F_1 through F_n, which are mixed together to form a new frequency. Each new frequency is mixed with the rf oscillator for transmission as one of the many hopping frequencies.

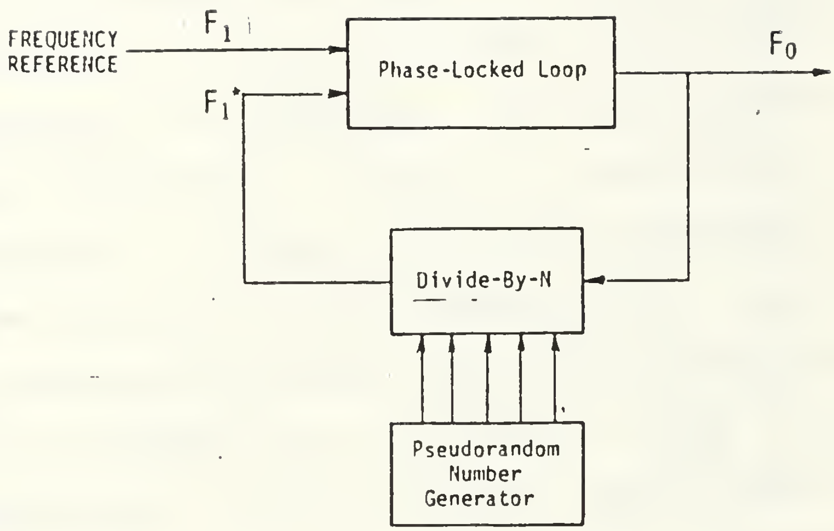


FIGURE 7.3 INDIRECT SIGNAL SYNTHESIS USING A PHASE-LOCKED LOOP

When the reference frequency, F_1, is at the same frequency as F_1^*, the voltage-controlled oscillator produces a constant output frequency, F_0. Since F_0 is divided by the programmable divider, F_0 is equal to N times F_1. By changing the divider ratio, N, many output frequencies are possible.

is shown in Figure (7.3) as the phase-locked loop output. The divide-by-N circuit divides this oscillator frequency by a selected number, N.

The phase-locked loop internally adjust its output frequency Fo so that F1* is the same frequency as the reference frequency F1. If the divide-by-N circuit output frequency is initially lower than the reference frequency F1, the oscillator output frequency is automatically increased until F1 and F1* are identical. When this occurs, the oscillator output frequency will become stable and remain at that frequency until the number, N, changes. When this number changes, the oscillator frequency is again automatically adjusted so that F1 and F1* are again identical.

B. PROPOSED USES OF PROGRAMMABLE FILTER

A wide field of applications exist in FH systems for the programmable GIC filter. The outstanding performance of the filter (including sensitivity and stability) and its high speed response to the different inputs (due to the use of CMOS integrated circuits) make it very exceptional in this field. The first proposed use is indicated, in Fig. (7.4). The figure illustrates a receiver of a frequency hopping demodulator. The received frequency hopped signal after heterodyned by the RF mixer passes through the GIC filter in a BP realization (at this application the topology control network does not need to exist since the BP realization is

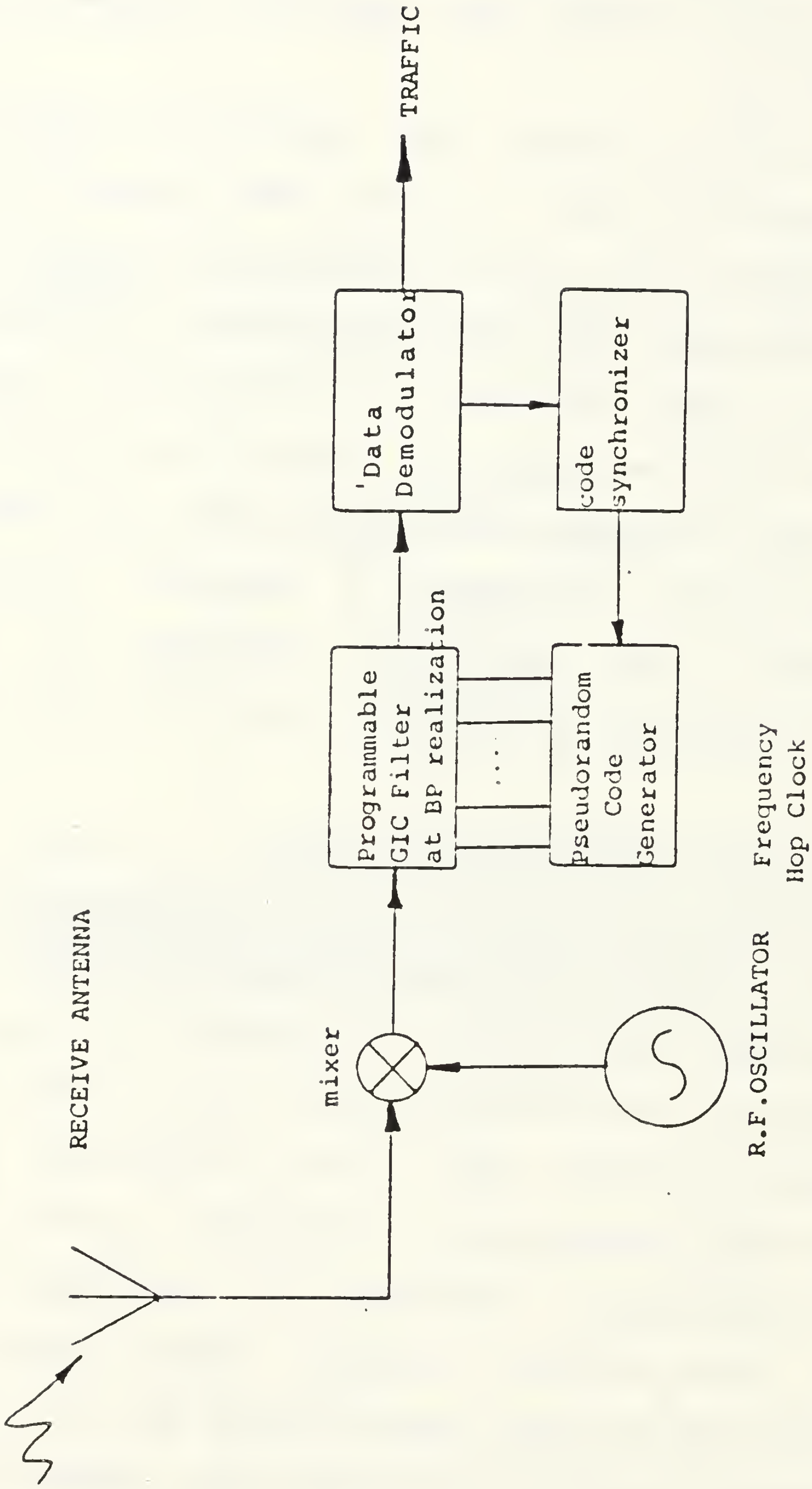


Fig. 7.4 - The Proposed Receiver of A Frequency Hopping System Using the Programmable GIC Filter

the only type to be used). The frequency shift of the filter is controlled by the synchronized pseudorandom code generator. This code generator also controls Q through some interfaced binary logic in order to correct the amplitude reduction which appears at high frequencies. The frequency shift problems can be easily corrected by the use of composate Op. Amp. Fig. (7.5) as it is extensively analyzed in Refs. [32], [34], and [35].

The programmable filter can also be used in the direct frequency synthesizer as it is illustrated at Fig. (7.2). At each frequency mixer output exists the need of a filter required to reject the unwanted frequencies. The frequencies to pass are not always the same, but they hop. the BP realization of the filter is used which center frequency can be controlled accordingly. The filter may require a short time for the signal to stabilize after it is selected. The time required for the filter to be stabilized at each new frequency may ultimately determine the maximum hopping rate of the direct frequency synthesizer.

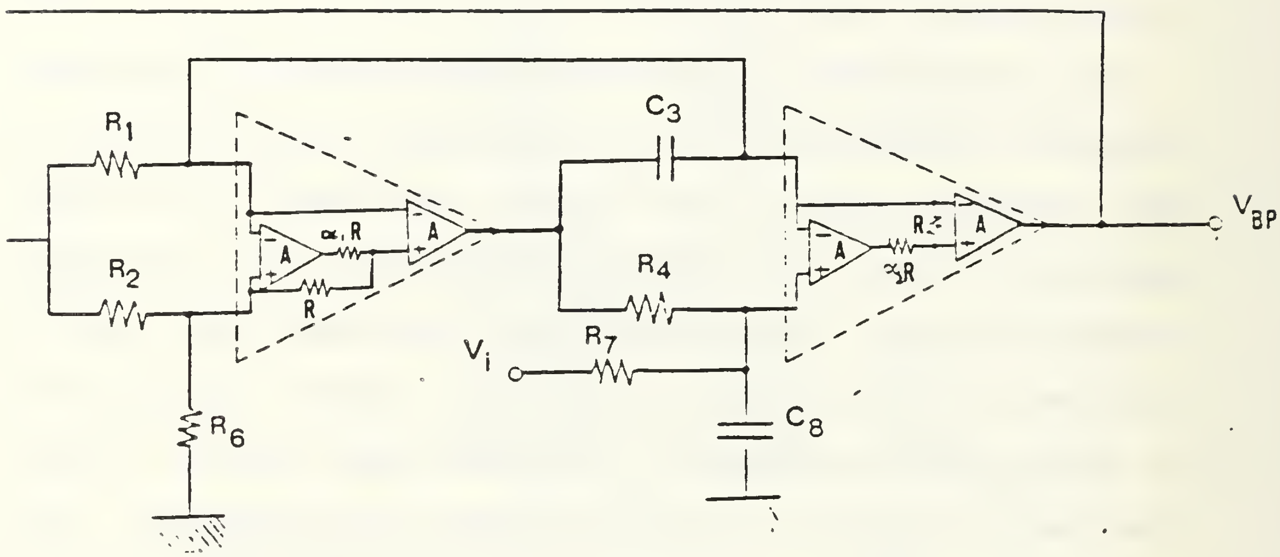


Fig. 7.5 Practical BP Filter Realization of the Composite GIC

VIII. CONCLUSION

The novel design described here has resulted in a universal programmable filter than can be digitally controlled to realize almost any practical filter specifications. This is done through the use of CMOS switches controlled by binary codes to program the order of the filter, the filter topology, the filter center frequency and selectivity. The design procedure required developing optimum switching arrangements for the minimum redundancy in components and least dependence of the filtering function on switching imperfections such as switches stray capacitances and non-zero and nonlinear switch-on resistance. The sensitivities of Qp, Wp are found to be low with respect to the passive and active elements variations. The experimental result show close agreement between theory and practice. Further, these results indicate that these realizations are insensitive to temperature and power supply variations. A wide field of applications exists for the programmable filter beside the one discussed in Chapter VII.

(1) Word recognition and speech synthesis;

(2) Music applications;

(3) Signal processing in communication;

(4) Adaptive balancing.

Further investigation is needed to develop a programmable switched capacitor realization that can allow frequency scaling by changing clock frequency. Work can be also extended for developing a wide bandwidth programmable filter using the composite operational amplifier technique proposed by [39]. Such implementation would lead to a very useful monolithic device at moderate cost.

The research has yielded a paper that was presented at the 19th Annual Asilomar Conference on circuits, systems and computers, Monterey, California; November, 1985 (Appendix C).

APPENDIX A

```
$JOB
      DIMENSION AT1S(100),PT1S(100),AT2S(100),PT2S(100)
     *,AT3S(100),PT3S(100),FAT1(100),FAT2(100),FAT3(100)
      DIMENSION AT1(100),PT1(100),AT2(100),PT2(100),AT3(100),
     * PT3(100),FPT1(100),FPT2(100),FPT3(100),
     * FTF1(100),FTF2(100),FTF3(100),FP1(100),FP2(100),FP3(100)
      COMPLEX T1S,T2S,T3S,Y1S,Y2S,Y3S,Y4S,Y5S,Y6S,Y7S,Y8S,DS,K1S,K2S,
     *K3S,K4S,K5S,K6S
      COMPLEX T1,T2,T3,Y1,Y2,Y3,Y4,Y5,Y6,Y7,Y8,D,K1,K2,K3,K4,K5,K6
      DIMENSION FR(100),3T1S(100),HT1S(100),3T2S(100),HT2S(100)
     *,BT3S(100),HT3S(100),FBT1(100),FBT2(100),FBT3(100)
      DIMENSION BT1(100),HT1(100),BT2(100),HT2(100),BT3(100)
     *,HT3(100),FHT1(100),FHT2(100),FHT3(100)
      COMPLEX C1S,C2S,C3S,G1S,G2S,G3S,G4S,G5S,G6S,G7S,G8S,DDS,X1S,X2S,
     *X3S,X4S,X5S,X6S
      COMPLEX C1,C2,C3,G1,G2,G3,G4,G5,G6,G7,G8,DD,X1,X2,X3,X4,X5,X6,3
      OMEGA = 0.
      DO 20 K = 1,100
      OMEGA = OMEGA + 125.*20.
      S = CMPLX( 0.0 , OMEGA )
C******************  DATA  ****************************************
      R1  = 1600.
      C1  = 200E-9
      RQ1 = 3200.
      R2  = 1600.
      C2  = 50E-9
      RQ2 = 3200.
      R3  = 1600.
      C3  = 12.5E-9
      RQ3 = 3200.
      R4  = 1600.
      C4  = 3.12E-9
      RQ4 = 3200.
      A3 = A4 = A3S = A4S = 1E5
      A1 = A2 = A1S = A2S = 1E5
      J = 4
      I = 4
      L = 4
      N = 4
C
C
C ****************SELECT THE TYPE OF THE FOURTH FILTER **************
      DO CASE L
          CASE
C          HPF
                 G1S = 1./R4
                 G2S = G1S
                 G3S = S*C4
                 G4S = G1S
                 G6S = G1S
                 G5S = 0.0
                 G7S = G3S
                 G8S = 1./RQ4
          CASE
C         LPF
                 G1S = 1./R4
                 G2S = S*C4
                 G3S = G2S+ 1./RQ4
                 G4S = G1S
                 G6S = 0.0
                 G5S = G1S
                 G7S = 0.0
                 G8S =G1S
          CASE
C         NOTCH
                 G1S = 1./R4
                 G2S = G1S
                 G3S = S*C4
                 G4S = G1S
                 G5S = G1S
                 G6S = 0.0
                 G7S = G3S
```

```
            G7S = G3S
            G8S = 1./RQ4
         IF NONE DO
C        BPF
            G1S = 1./R4
            G2S = G1S
            G3S = S*C4
            G4S = G1S
            G5S = 0.0
            G6S = G1S
            G7S = 1./RQ4
            G8S =S*C4
         END CASE
C
C*************** SELECT THE TYPE OF THE THIRD FILTER ****************
         DO CASE N
         CASE
C         HPF
            G1  = 1./R3
            G2  = G1
            G3  = S*C3
            G4  = G1
            G6  = G1
            G5  = 0.0
            G7  = G3
            G8  = 1./RQ3
         CASE
C        LPF
            G1  = 1./R3
            G2  = S*C3
            G3  = G2 + 1./RQ3
            G4  = G1
            G6  = 0.0
            G5  = G1
            G7  = 0.0
            G8  = G1
         CASE
C        NOTCH
            G1 = 1./R3
            G2 = G1
            G3 = S*C3
            G4 = G1
            G5 = G1
            G6 = 0.
            G7 = G3
            G8 = 1./RQ3
         IF NONE DO
C        BPF
            G1  = 1./R3
            G2  = G1
            G3  = S*C3
            G4  = G1
            G5  = 0.0
            G6  = G1
            G7  = 1./RQ3
            G8  =S*C3
         END CASE
C ***************SELECT THE TYPE OF THE SECOND FILTER **************
      DO CASE I
         CASE
C         HPF
            Y1S = 1./R2
            Y2S = Y1S
            Y3S = S*C2
            Y4S = Y1S
            Y6S = Y1S
            Y5S = 0.0
            Y7S = Y3S
            Y8S = 1./RQ2
         CASE
C        LPF
            Y1S = 1./R2
```

```
                Y2S  =  S*C2
                Y3S  =  Y2S+  1./RQ2
                Y4S  =  Y1S
                Y6S  =  0.0
                Y5S  =  Y1S
                Y7S  =  0.0
                Y8S  =Y1S
          CASE
C         NOTCH
                Y1S  =  1./R2
                Y2S  =  Y1S
                Y3S  =  S*C2
                Y4S  =  Y1S
                Y5S  =  Y1S
                Y6S  =  0.0
                Y7S  =  Y3S
                Y8S  =  1./RQ2
          IF NONE DO
C         BPF
                Y1S  =  1./R2
                Y2S  =  Y1S
                Y3S  =  S*C2
                Y4S  =  Y1S
                Y5S  =  0.0
                Y6S  =  Y1S
                Y7S  =  1./RQ2
                Y8S  =S*C2
          END CASE
C
C**************** SELECT THE TYPE OF THE FIRST FILTER ****************
          DO CASE J
          CASE
C          HPF
                Y1   =  1./R1
                Y2   =  Y1
                Y3   =  S*C1
                Y4   =  Y1
                Y6   =  Y1
                Y5   =  0.0
                Y7   =  Y3
                Y8   =  1./RQ1
          CASE
C         LPF
                Y1   =  1./R1
                Y2   =  S*C1
                Y3   =  Y2 +  1./RQ1
                Y4   =  Y1
                Y6   =  0.0
                Y5   =  Y1
                Y7   =  0.0
                Y8   =  Y1
          CASE
C         NOTCH
                Y1  =  1./R1
                Y2  =  Y1
                Y3  =  S*C1
                Y4  =  Y1
                Y5  =  Y1
                Y6  =  0.
                Y7  =  Y3
                Y3  =  1./RQ1
          IF NONE DO
C         BPF
                Y1   =  1./R1
                Y2   =  Y1
                Y3   =  S*C1
                Y4   =  Y1
                Y5   =  0.0
                Y6   =  Y1
                Y7   =  1./RQ1
                Y8   =S*C1
          END CASE
```

```
C**********************************************************************
C
      FR(K) = OMEGA/6.28
      S = CMPLX(0.0,OMEGA)
      K1S = Y1S*(1.+ 1./A2S)
      K2S = (Y2S + Y5S + Y6S)
      K3S = (Y4S + Y7S + Y8S)
      K4S = Y4S + Y8S
      K5S = Y1S + Y3S
      K6S = Y5S + Y6S
C
      K1 = Y1*(1.+ 1./A2)
      K2 = (Y2 + Y5 + Y6)
      K3 = (Y4 + Y7 + Y8)
      K4 = Y4 + Y8
      K5 = Y1 + Y3
      K6 = Y5 + Y6
C
      X1S = G1S*(1.+ 1./A4S)
      X2S = (G2S + G5S + G6S)
      X3S = (G4S + G7S + G8S)
      X4S = G4S + G8S
      X5S = G1S + G3S
      X6S = G5S + G6S
C
      X1 = G1*(1.+ 1./A4)
      X2 = (G2 + G5 + G6)
      X3 = (G4 + G7 + G8)
      X4 = G4 + G8
      X5 = G1 + G3
      X6 = G5 + G6
C
C
      DS =(K1S*K2S*K3S)/A1S + Y1S*Y4S*K6S +Y3S*K2S*K3S/(A1S*A2S)+
     *   Y2S*Y3S*Y4S/A2S+(Y2S/Y1S)*K1S*Y3S*(Y7S+Y8S) + Y3S*K6S*K3S/A2S
     T1S=(Y1S*Y4S*Y5S +Y3S*Y5S*(Y7S-Y8S) +Y3S*Y7S*(Y2S+Y6S)+(Y7S*
     *    (Y1S+Y3S)*K2S)/A1S)/DS
     T2S=(Y1S*Y5S*K4S -Y1S*Y7S*Y6S +Y3S*Y5S*K4S/A2S -Y3S*Y6S*Y7S/A2S
     *      +Y5S*Y7S*K5S/A1S + Y2S*Y3S*Y7S )/DS
     T3S=(K1S*Y7S*K2S/A1S +K1S*Y4S*Y5S +Y3S*Y7S*K2S/A1S*A2S +Y3S*Y4S
     * *Y5S/A2S + (Y1S/Y2S)*K1S*Y3S*Y7S + K6S*Y3S*Y7S/A2S)/DS
C
      D =(K1*K2*K3)/A1 + Y1*Y4*K6 +Y3*K2*K3/(A1*A2)+ Y2*Y3*Y4/A2 +
     *     (Y2/Y1)*K1*Y3*(Y7+Y8) + Y3*K6*K3/A2
     T1=(Y1*Y4*Y5 +Y3*Y5*(Y7-Y8) +Y3*Y7*(Y2+Y6)+(Y7*(Y1+Y3)*K2)/A1)/D
     T2=(Y1*Y5*K4 -Y1*Y7*Y6 +Y3*Y5*K4/A2 -Y3*Y6*Y7/A2 +Y5*Y7*K5/A2 +
     *    Y2*Y3*Y7 )/D
     T3=(K1*Y7*K2/A1 +K1*Y4*Y5 +Y3*Y7*K2/A1*A2 +Y3*Y4*Y5/A2 +(Y2/Y1)*
     *      K1*Y3*Y7 +K6*Y3*Y7/A2)/D
C
C
      DDS =(X1S*X2S*X3S)/A3S + G1S*G4S*X6S +G3S*X2S*X3S/(A3S*A4S)+
     *   G2S*G3S*G4S/A4S+(G2S/G1S)*X1S*G3S*(G7S+G8S) + G3S*X6S*X3S/A4S
     C1S=(G1S*G4S*G5S +G3S*G5S*(G7S-G8S) +G3S*G7S*(G2S+G6S)+(G7S*
     *    (G1S+G3S)*X2S)/A3S)/DDS
     C2S=(G1S*G5S*X4S -G1S*G7S*G6S +G3S*G5S*X4S/A4S -G3S*G6S*G7S/A4S
     *      +G5S*G7S*X5S/A3S + G2S*G3S*G7S )/DDS
     C3S=(X1S*G7S*X2S/A3S +X1S*G4S*G5S +G3S*G7S*X2S/A3S*A4S +G3S*G4S
     * *G5S/A4S + (G1S/G2S)*X1S*G3S*G7S + X6S*G3S*G7S/A4S)/DDS
C
      DD=(X1*X2*X3)/A3 + G1*G4*X6 +G3*X2*X3/(A3*A4)+ G2*G3*G4/A4 +
     *     (G2/G1)*X1*G3*(G7+G8) + G3*X6*X3/A4
     C1=(G1*G4*G5+G3*G5*(G7-G8)+G3*G7*(G2+G6)+(G7*(G1+G3)*X2)/A3)/DD
     C2=(G1*G5*X4-G1*G7*G6 +G3*G5*X4/A4 -G3*G6*G7/A4 +G5*G7*X5/A4 +
     *    G2*G3*G7 )/DD
     C3=(X1*G7*X2/A3 +X1*G4*G5 +G3*G7*X2/A3*A4 +G3*G4*G5/A4 +(G2/G1)*
     *      X1*G3*G7 +X6*G3*G7/A4)/DD
C
C
      AT1S(K) = 20.*ALOG10(CABS(T1S))
      AT2S(K) = 20.*ALOG10(CABS(T2S))
      AT3S(K) = 20.*ALOG10(CABS(T3S))
```

```
C
      ART1S = REAL(T1S)
      ART2S = REAL(T2S)
      ART3S = REAL(T3S)
C
      AIT1S = AIMAG(T1S)
      AIT2S = AIMAG(T2S)
      AIT3S = AIMAG(T3S)
C
      PT1S(K) = ATAN2(AIT1S , ART1S)* 57.325
      PT2S(K) = ATAN2(AIT2S , ART2S)* 57.325
      PT3S(K) = ATAN2(AIT3S , ART3S)* 57.325
C
      AT1(K) = 20.*ALOG10(CABS(T1))
      AT2(K) = 20.*ALOG10(CABS(T2))
      AT3(K) = 20.*ALOG10(CABS(T3))
C
      ART1 = REAL(T1)
      ART2 = REAL(T2)
      ART3 = REAL(T3)
C
      AIT1 = AIMAG(T1)
      AIT2 = AIMAG(T2)
      AIT3 = AIMAG(T3)
C
      PT1(K) = ATAN2(AIT1 , ART1)* 57.325
      PT2(K) = ATAN2(AIT2 , ART2)* 57.325
      PT3(K) = ATAN2(AIT3 , ART3)* 57.325
C
      FAT1(K) = AT1(K) + AT1S(K)
      FAT2(K) = AT2(K) + AT2S(K)
      FAT3(K) = AT3(K) + AT3S(K)
C
      FPT1(K) = PT1(K) + PT1S(K)
      FPT2(K) = PT2(K) + PT2S(K)
      FPT3(K) = PT3(K) + PT3S(K)
C
      BT1S(K) = 20.*ALOG10(CABS(C1S))
      BT2S(K) = 20.*ALOG10(CABS(C2S))
      BT3S(K) = 20.*ALOG10(CABS(C3S))
C
      ART1S = REAL(C1S)
      ART2S = REAL(C2S)
      ART3S = REAL(C3S)
C
      AIT1S = AIMAG(C1S)
      AIT2S = AIMAG(C2S)
      AIT3S = AIMAG(C3S)
C
      HT1S(K) = ATAN2(AIT1S , ART1S)* 57.325
      HT2S(K) = ATAN2(AIT2S , ART2S)* 57.325
      HT3S(K) = ATAN2(AIT3S , ART3S)* 57.325
C
      BT1(K) = 20.*ALOG10(CABS(C1))
      BT2(K) = 20.*ALOG10(CABS(C2))
      BT3(K) = 20.*ALOG10(CABS(C3))
C
      ART1 = REAL(C1)
      ART2 = REAL(C2)
      ART3 = REAL(C3)
C
      AIT1 = AIMAG(C1)
      AIT2 = AIMAG(C2)
      AIT3 = AIMAG(C3)
C
      HT1(K) = ATAN2(AIT1 , ART1)* 57.325
      HT2(K) = ATAN2(AIT2 , ART2)* 57.325
      HT3(K) = ATAN2(AIT3 , ART3)* 57.325
C
      FBT1(K) = BT1(K) + BT1S(K)
      FBT2(K) = BT2(K) + BT2S(K)
      FBT3(K) = BT3(K) + BT3S(K)
```

```
C
          FHT1(K) = HT1(K)  + HT1S(K)
          FHT2(K) = HT2(K)  + HT2S(K)
          FHT3(K) = HT3(K)  + HT3S(K)
C
          FTF1(K) = AT1(K) +  BT1(K) + BT1S(K)+ AT1S(K)
          FTF2(K) = FAT2(K) + FBT2(K)
          FTF3(K) = FAT3(K) + FBT3(K)
C
          FP1(K) = FHT1(K) + FPT1(K)
          FP2(K) = FHT2(K) + FPT2(K)
          FP3(K) = FHT3(K) + FPT3(K)
C
          WRITE(6,66) FR(K),FAT1(K),FBT1(K)
     *                                          ,FTF1(K),HT1(K),FHT1(K)
C    *    FHT3(K)
20      CONTINUE
66      FORMAT(6(1X,F9.3))
        STOP
        END
$ENTRY
```

APPENDIX B

```
$JOB
      DIMENSION AT1S(100),PT1S(100),AT2S(100),PT2S(100)
     *,AT3S(100),PT3S(100),FAT1(100),FAT2(100),FAT3(100)
      DIMENSION AT1(100),PT1(100),AT2(100),PT2(100),AT3(100),
     * PT3(100),FPT1(100),FPT2(100),FPT3(100),
     * FTF1(100),FTF2(100),FTF3(100),FP1(100),FP2(100),FP3(100)
      COMPLEX T1S,T2S,T3S,Y1S,Y2S,Y3S,Y4S,Y5S,Y6S,Y7S,Y8S,DS,K1S,K2S,
     *K3S,K4S,K5S,K6S
      COMPLEX T1,T2,T3,Y1,Y2,Y3,Y4,Y5,Y6,Y7,Y8,D,K1,K2,K3,K4,K5,K6
      DIMENSION FR(100),BT1S(100),HT1S(100),BT2S(100),HT2S(100)
     *,BT3S(100),HT3S(100),FBT1(100),FBT2(100),FBT3(100)
      DIMENSION BT1(100),HT1(100),BT2(100),HT2(100),BT3(100)
     *,HT3(100),FHT1(100),FHT2(100),FHT3(100)
      COMPLEX C1S,C2S,C3S,G1S,G2S,G3S,G4S,G5S,G6S,G7S,G8S,DDS,X1S,X2S,
     *X3S,X4S,X5S,X6S
      COMPLEX C1,C2,C3,G1,G2,G3,G4,G5,G6,G7,G8,DD,X1,X2,X3,X4,X5,X6,S
      OMEGA = 0.
      DO 20 K = 1,100
      OMEGA = OMEGA + 125.*20.
      S = CMPLX( 0.0 , OMEGA )
C******************  DATA ******************************************
      R1  = 1600.
      C1  = 200E-9
      RQ1 = 3200.
      R2  = 1600.
      C2  = 50E-9
      RQ2 = 3200.
      R3  = 1600.
      C3  = 12.5E-9
      RQ3 = 3200.
      R4  = 1600.
      C4  = 3.12E-9
      RQ4 = 3200.
W1 = W1S = "SET VALUE "
      A3 = A4 = A3S = A4S = W1/S
      A1 = A2 = A1S = A2S = W1S/S
      J = 4
      I = 4
      L = 4
      N = 4
C
C
C ****************SELECT THE TYPE OF THE FOURTH FILTER **************
      DO CASE L
         CASE
C          HPF
                G1S = 1./R4
                G2S = G1S
                G3S = S*C4
                G4S = G1S
                G6S = G1S
                G5S = 0.0
                G7S = G3S
                G8S = 1./RQ4
         CASE
C        LPF
                G1S = 1./R4
                G2S = S*C4
                G3S = G2S+ 1./RQ4
                G4S = G1S
                G6S = 0.0
                G5S = G1S
                G7S = 0.0
                G8S =G1S
         CASE
C        NOTCH
                G1S = 1./R4
                G2S = G1S
                G3S = S*C4
                G4S = G1S
                G5S = G1S
                G6S = 0.0
```

```
            G7S = G3S
            G8S = 1./RQ4
        IF NONE DO
C       BPF
            G1S = 1./R4
            G2S = G1S
            G3S = S*C4
            G4S = G1S
            G5S = 0.0
            G6S = G1S
            G7S = 1./RQ4
            G8S =S*C4
        END CASE
C
C**************** SELECT THE TYPE OF THE THIRD FILTER *******************
        DO CASE N
        CASE
C        HPF
            G1  = 1./R3
            G2  = G1
            G3  = S*C3
            G4  = G1
            G6  = G1
            G5  = 0.0
            G7  = G3
            G8  = 1./RQ3
        CASE
C       LPF
            G1  = 1./R3
            G2  = S*C3
            G3  = G2 + 1./RQ3
            G4  = G1
            G6  = 0.0
            G5  = G1
            G7  = 0.0
            G8  = G1
        CASE
C       NOTCH
            G1 = 1./R3
            G2 = G1
            G3 = S*C3
            G4 = G1
            G5 = G1
            G6 = 0.
            G7 = G3
            G8 = 1./RQ3
        IF NONE DO
C       BPF
            G1  = 1./R3
            G2  = G1
            G3  = S*C3
            G4  = G1
            G5  = 0.0
            G6  = G1
            G7  = 1./RQ3
            G8  =S*C3
        END CASE
C ****************SELECT THE TYPE OF THE SECOND FILTER ***************
      DO CASE I
        CASE
C        HPF
            Y1S = 1./R2
            Y2S = Y1S
            Y3S = S*C2
            Y4S = Y1S
            Y6S = Y1S
            Y5S = 0.0
            Y7S = Y3S
            Y8S = 1./RQ2
        CASE
C       LPF
            Y1S = 1./R2
```

```
                Y2S = S*C2
                Y3S = Y2S+ 1./RQ2
                Y4S = Y1S
                Y6S = 0.0
                Y5S = Y1S
                Y7S = 0.0
                Y8S =Y1S
         CASE
C        NOTCH
                Y1S = 1./R2
                Y2S = Y1S
                Y3S = S*C2
                Y4S = Y1S
                Y5S = Y1S
                Y6S = 0.0
                Y7S = Y3S
                Y8S = 1./RQ2
         IF NONE DO
C        BPF
                Y1S = 1./R2
                Y2S = Y1S
                Y3S = S*C2
                Y4S = Y1S
                Y5S = 0.0
                Y6S = Y1S
                Y7S = 1./RQ2
                Y8S =S*C2
         END CASE
C
C**************** SELECT THE TYPE OF THE FIRST FILTER *****************
         DO CASE J
         CASE
C         HPF
                Y1  = 1./R1
                Y2  = Y1
                Y3  = S*C1
                Y4  = Y1
                Y6  = Y1
                Y5  = 0.0
                Y7  = Y3
                Y8  = 1./RQ1
         CASE
C        LPF
                Y1  = 1./R1
                Y2  = S*C1
                Y3  = Y2 + 1./RQ1
                Y4  = Y1
                Y6  = 0.0
                Y5  = Y1
                Y7  = 0.0
                Y8  = Y1
         CASE
C        NOTCH
                Y1 = 1./R1
                Y2 = Y1
                Y3 = S*C1
                Y4 = Y1
                Y5 = Y1
                Y6 = 0.
                Y7 = Y3
                Y3 = 1./RQ1
         IF NONE DO
C        BPF
                Y1  = 1./R1
                Y2  = Y1
                Y3  = S*C1
                Y4  = Y1
                Y5  = 0.0
                Y6  = Y1
                Y7  = 1./RQ1
                Y8  =S*C1
         END CASE
```

```
C*******************************************************************************
C
      FR(K) = OMEGA/6.28
      S = CMPLX(0.0,OMEGA)
      K1S = Y1S*(1.+ 1./A2S)
      K2S = (Y2S + Y5S + Y6S)
      K3S = (Y4S + Y7S + Y8S)
      K4S = Y4S + Y8S
      K5S = Y1S + Y3S
      K6S = Y5S + Y6S
C
      K1 = Y1*(1.+ 1./A2)
      K2 = (Y2 + Y5 + Y6)
      K3 = (Y4 + Y7 + Y8)
      K4 = Y4 + Y8
      K5 = Y1 + Y3
      K6 = Y5 + Y6
C
      X1S = G1S*(1.+ 1./A4S)
      X2S = (G2S + G5S + G6S)
      X3S = (G4S + G7S + G8S)
      X4S = G4S + G8S
      X5S = G1S + G3S
      X6S = G5S + G6S
C
      X1 = G1*(1.+ 1./A4)
      X2 = (G2 + G5 + G6)
      X3 = (G4 + G7 + G8)
      X4 = G4 + G8
      X5 = G1 + G3
      X6 = G5 + G6
C
C
      DS =(K1S*K2S*K3S)/A1S + Y1S*Y4S*K6S +Y3S*K2S*K3S/(A1S*A2S)+
     *  Y2S*Y3S*Y4S/A2S+(Y2S/Y1S)*K1S*Y3S*(Y7S+Y8S) + Y3S*K6S*K3S/A2
     T1S=(Y1S*Y4S*Y5S +Y3S*Y5S*(Y7S-Y8S) +Y3S*Y7S*(Y2S+Y6S)+(Y7S*
     *   (Y1S+Y3S)*K2S)/A1S)/DS
     T2S=(Y1S*Y5S*K4S -Y1S*Y7S*Y6S +Y3S*Y5S*K4S/A2S -Y3S*Y6S*Y7S/A2S
     *   +Y5S*Y7S*K5S/A1S + Y2S*Y3S*Y7S )/DS
     T3S=(K1S*Y7S*K2S/A1S +K1S*Y4S*Y5S +Y3S*Y7S*K2S/A1S*A2S +Y3S*Y4S
     * *Y5S/A2S + (Y1S/Y2S)*K1S*Y3S*Y7S + K6S*Y3S*Y7S/A2S)/DS
C
      D =(K1*K2*K3)/A1 + Y1*Y4*K6 +Y3*K2*K3/(A1*A2)+ Y2*Y3*Y4/A2 +
     *   (Y2/Y1)*K1*Y3*(Y7+Y8) + Y3*K6*K3/A2
     T1=(Y1*Y4*Y5 +Y3*Y5*(Y7-Y8) +Y3*Y7*(Y2+Y6)+(Y7*(Y1+Y3)*K2)/A1)/
     T2=(Y1*Y5*K4 -Y1*Y7*Y6 +Y3*Y5*K4/A2 -Y3*Y6*Y7/A2 +Y5*Y7*K5/A2 +
     *   Y2*Y3*Y7 )/D
     T3=(K1*Y7*K2/A1 +K1*Y4*Y5 +Y3*Y7*K2/A1*A2 +Y3*Y4*Y5/A2 +(Y2/Y1)
     *       K1*Y3*Y7 +K6*Y3*Y7/A2)/D
C
C
      DDS =(X1S*X2S*X3S)/A3S + G1S*G4S*X6S +G3S*X2S*X3S/(A3S*A4S)+
     *  G2S*G3S*G4S/A4S+(G2S/G1S)*X1S*G3S*(G7S+G8S) + G3S*X6S*X3S/A4
     C1S=(G1S*G4S*G5S +G3S*G5S*(G7S-G8S) +G3S*G7S*(G2S+G6S)+(G7S*
     *   (G1S+G3S)*X2S)/A3S)/DDS
     C2S=(G1S*G5S*X4S -G1S*G7S*G6S +G3S*G5S*X4S/A4S -G3S*G6S*G7S/A4S
     *   +G5S*G7S*X5S/A3S + G2S*G3S*G7S )/DDS
     C3S=(X1S*G7S*X2S/A3S +X1S*G4S*G5S +G3S*G7S*X2S/A3S*A4S +G3S*G4S
     * *G5S/A4S + (G1S/G2S)*X1S*G3S*G7S + X6S*G3S*G7S/A4S)/DDS
C
      DD=(X1*X2*X3)/A3 + G1*G4*X6 +G3*X2*X3/(A3*A4)+ G2*G3*G4/A4 +
     *   (G2/G1)*X1*G3*(G7+G8) + G3*X6*X3/A4
     C1=(G1*G4*G5+G3*G5*(G7-G8)+G3*G7*(G2+G6)+(G7*(G1+G3)*X2)/A3)/DD
     C2=(G1*G5*X4-G1*G7*G6 +G3*G5*X4/A4 -G3*G6*G7/A4 +G5*G7*X5/A4 +
     *   G2*G3*G7 )/DD
     C3=(X1*G7*X2/A3 +X1*G4*G5 +G3*G7*X2/A3*A4 +G3*G4*G5/A4 +(G2/G1)
     *       X1*G3*G7 +X6*G3*G7/A4)/DD
C
C
      AT1S(K) = 20.*ALOG10(CABS(T1S))
      AT2S(K) = 20.*ALOG10(CABS(T2S))
      AT3S(K) = 20.*ALOG10(CABS(T3S))
```

```
C
      ARTIS = REAL(T1S)
      ART2S = REAL(T2S)
      ART3S = REAL(T3S)
C
      AITIS = AIMAG(T1S)
      AIT2S = AIMAG(T2S)
      AIT3S = AIMAG(T3S)
C
      PT1S(K) = ATAN2(AIT1S , ART1S)* 57.325
      PT2S(K) = ATAN2(AIT2S , ART2S)* 57.325
      PT3S(K) = ATAN2(AIT3S , ART3S)* 57.325
C
      AT1(K) = 20.*ALOG10(CABS(T1))
      AT2(K) = 20.*ALOG10(CABS(T2))
      AT3(K) = 20.*ALOG10(CABS(T3))
C
      ART1 = REAL(T1)
      ART2 = REAL(T2)
      ART3 = REAL(T3)
C
      AIT1 = AIMAG(T1)
      AIT2 = AIMAG(T2)
      AIT3 = AIMAG(T3)
C
      PT1(K) = ATAN2(AIT1 , ART1)* 57.325
      PT2(K) = ATAN2(AIT2 , ART2)* 57.325
      PT3(K) = ATAN2(AIT3 , ART3)* 57.325
C
      FAT1(K) = AT1(K) + AT1S(K)
      FAT2(K) = AT2(K) + AT2S(K)
      FAT3(K) = AT3(K) + AT3S(K)
C
      FPT1(K) = PT1(K) + PT1S(K)
      FPT2(K) = PT2(K) + PT2S(K)
      FPT3(K) = PT3(K) + PT3S(K)
C
      BT1S(K) = 20.*ALOG10(CABS(C1S))
      BT2S(K) = 20.*ALOG10(CABS(C2S))
      BT3S(K) = 20.*ALOG10(CABS(C3S))
C
      ART1S = REAL(C1S)
      ART2S = REAL(C2S)
      ART3S = REAL(C3S)
C
      AIT1S = AIMAG(C1S)
      AIT2S = AIMAG(C2S)
      AIT3S = AIMAG(C3S)
C
      HT1S(K) = ATAN2(AIT1S , ART1S)* 57.325
      HT2S(K) = ATAN2(AIT2S , ART2S)* 57.325
      HT3S(K) = ATAN2(AIT3S , ART3S)* 57.325
C
      BT1(K) = 20.*ALOG10(CABS(C1))
      BT2(K) = 20.*ALOG10(CABS(C2))
      BT3(K) = 20.*ALOG10(CABS(C3))
C
      ART1 = REAL(C1)
      ART2 = REAL(C2)
      ART3 = REAL(C3)
C
      AIT1 = AIMAG(C1)
      AIT2 = AIMAG(C2)
      AIT3 = AIMAG(C3)
C
      HT1(K) = ATAN2(AIT1 , ART1)* 57.325
      HT2(K) = ATAN2(AIT2 , ART2)* 57.325
      HT3(K) = ATAN2(AIT3 , ART3)* 57.325
C
      FBT1(K) = BT1(K) + BT1S(K)
      FBT2(K) = BT2(K) + BT2S(K)
      FBT3(K) = BT3(K) + BT3S(K)
```

```
C
          FHT1(K) = HT1(K) + HT1S(K)
          FHT2(K) = HT2(K) + HT2S(K)
          FHT3(K) = HT3(K) + HT3S(K)
C
          FTF1(K) = AT1(K) +  BT1(K) + BT1S(K)+ AT1S(K)
          FTF2(K) = FAT2(K) + FBT2(K)
          FTF3(K) = FAT3(K) + FBT3(K)
C
          FP1(K) = FHT1(K) + FPT1(K)
          FP2(K) = FHT2(K) + FPT2(K)
          FP3(K) = FHT3(K) + FPT3(K)
C
          WRITE(6,66) FR(K),FAT1(K),FBT1(K)
     *                                          ,FTF1(K),HT1(K),FHT1(K)
C    *    FHT3(K)
20      CONTINUE
66      FORMAT(6(1X,F9.3))
        STOP
        END
$ENTRY
```

APPENDIX C

DIGITALLY CONTROLLED "PROGRAMMABLE" ACTIVE FILTERS

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ABSTRACT

In this contribution a general purpose digitally controlled analog filter is presented. The novel design is a cascade of second order sections that are individually programmed to achieve any filtering topologies. Two-binary words are used to control the pole frequency ω_p and selectivity Q_p of each section independently. Each second order section is a Generalized-Immittance converter (GIC) biquads which are known for their high stability and low active and passive sensitivity. CMOS switches are used to electronically relocate the minimum number of passive elements to achieve function programmability. Switches are also used to select the number of cascaded sections to realize higher order transfer functions.

1. INTRODUCTION

The availability of an analog filter with digitally controlled "programmable" coefficients has been the goal of many researchers due to its several attractions. One possibility of a compact, versatile analog filter under remote control opens up many novel and independent application areas. Also, when a programmable filter is combined with a permanent reference memory which is user-programmable, this would form an economical and versatile device for dedicated stand-alone applications. The need for such a device was motivated by advancement in film and semiconductor technologies as well as the continuous upgrading of systems specifications to take advantage of the available technologies to the limits.

Linear analog filtering finds many applications, such as speech processing (recognition or synthesis), geology, instrumentation, communications, process control, adaptive balancing, etc. There has been much emphasis on performing the filtering function digitally, largely because of the ease of varying and optimizing the transfer function. However, for many reasons, such as cost, size, signal processing complexity, and bandwidth, it would be desirable to perform the filter function with linear components yet retain the flexability of varying the filter parameters digitally.

Recently, the advantages of combining linear components (operational amplifiers (OAS), resistors and capacitors) and nonlinear elements (switches) have been demonstrated using switched capacitor techniques [1-3]. In this contribution, we are presenting the results of realizing a continuous active device using linear elements and switches controlled by digital signals to achieve a fully programmable filter [4]. Several programming features of the proposed filter are reported. The first feature is the ability of the network to realize the most common filtering functions (function programmability) namely: Low Pass (LP), Band Pass (BP), High Pass (HP), All Pass (AP) and Notch(N) functions, using the minimal set of elements. The second feature is the ability of the network to program (independently) the key parameters of the filtering function chosen (parameter programmability) namely: the pole resonent frequency (ωp) and selectivity (Qp). Finally the ability to program the network to cascade several sections to achieve higher order filter. All of the above programmability features are performed independently to realize a universal filtering network.

2. DESIGN ANALYSIS OF THE PROPOSED FILTERS

The basic active network considered here as the heart of the programmable filter is the second order Generalized Immittance Converter (GIC) structure [5], Fig. 1, whose superior performance was established in the literature [6]. The general transfer function realized T(s) is given by:

$$T(s) = N(s)/D(s) = (a_0+a_1s+as^2)/(b_0+b_1s+b_2s^2) \quad (1)$$

The GIC transfer functions of Fig. 1 assuming non-ideal OAs are given by

$$\begin{bmatrix} Y_1\left[1+\frac{s}{\omega_2}\right]+\frac{Y_2}{\omega_2}s & Y_3 & -Y_3-Y_1 \\ Y_2\left[1+\frac{s}{\omega_2}\right]+\frac{s}{\omega_2}\left[Y_5+Y_6\right] & -\frac{s}{\omega_1}\left[Y_2+Y_5+Y_6\right] & -Y_2-Y_5-Y_6 \\ 0 & Y_4 & -Y_4-Y_7-Y_8 \end{bmatrix} \begin{bmatrix} T_1(s) \\ T_2(s) \\ T_3(s) \end{bmatrix} = \begin{bmatrix} 0 \\ -Y_5 \\ -Y_7 \end{bmatrix} \quad (2)$$

<u>Stability and sensitivity analysis:</u>

An important criterion of a realization is its sensitivity to element variations. The GIC sensitivity analysis has shown to be as good or better than all competitive second order networks [6]. While the GIC stability can easily be demonstrated since in all the transfer functions (2), the coefficients of D(s) are seen to remain positive for any OA mismatch. This is due to the absence of negative terms in D(s). Therefore the zeros of D(s) will remain in the left-half s-plane and low frequency unstable modes cannot arise during activation.

Function programmability:

The objective of this research was to develop a device that is capable of realizing the following transfer functions: LP where $T(s) = K/D(s)$, BP where $T(s) = KS/D(s)$, HP where $T(s) = KS^2/D(s)$, AP where $T(s)=[s^2-s(\omega_p/Q_p)+\omega_p^2]/D(s)$ and N where $T(s)=(s^2+\omega_z^2)/D(s)$. By optimizing the design of the filter, it was found that all of the above functions can be realized by the second order GIC section using four resistors, two capacitors and two OAs as shown in Table 1. These passive elements are connected to different nodes to achieve the various realizations. A set of CMOS bilateral switches controlled by digital binary word, are used to relocate the same elements in different ways to achieve the desired filtering functions according to Fig. 2. The truth table of the switches control logic is shown in Table 2, while Fig. 3 illustrates the corresponding minimized CMOS logic circuit used for passive elements relocation.

Paramater programmability:

While four of the resistors are equal and of value R each, the fifth resistor is the Q_p determining resistor and of value $R_q=RQ_p$. The two capacitors are equal and of value $c=1/(\omega_p R)$ each. Two equal banks of capacitors are used to control ωp. Each bank contains n binary-weighted capacitors connected in series with n CMOS switches as shown in Fig. 4. Using n bit binary word to control the switches, 2^n different values of c can be obtained that corresponds to 2^n different values of ωp. Using a similar technique, the value of R_q can be controlled by an m bit digital word that yields 2^m different values of Q_p, as illustrated in Fig. 5. Thus, full independent control of the pole pair ω_p and Q_p are achieved by programming the digital words controlling the switches to obtain the corresponding c and R_q. The complete second order programmable filter is shown in Fig.6.a where the function programmability as well as the parameter programmability are demonstrated.

Higher order programmability:

Active filters design procedure can be classified as direct or cascade. In direct synthesis procedures the transfer function is realized as a single section [7]. In cascade synthesis procedures a high order transfer function is expressed as a product of first and second order transfer functions and each of these is realized independently. The overall network is obtained by cascading the individual sections. The cascade method of synthesis offers two practical advantages (a) simple network tuning (b) a few number of universal sections can be designed which can realize a multitude of network specifications.

The second order GIC network structure lends itself to the cascade synthesis procedure since it does not require additional isolating amplifiers. Fig.6.b shows a block diagram of a programmable higher order filter that utilizes the second procedure by cascading 2 or more sections of the filter network shown in Fig. 6.a. The result is a high order fully programmable general purpose filter, that can be tailored to match almost any proposed specification.

3. COMPUTER SIMULATIONS AND EXPERIMENTAL VERIFICATIONS:

Fig. 7 shows differenct computer simulation outputs of the programmable filter. The plots simulate the filter responses assuming ideal OAs with infinite Gain Bandwidth Products (GBWP), as well as practical filter responses assuming OA's finite GBWPs of 1 MHz as of that of the LM741 OA. A single pole OA model was utilized to approximate the filter transfer functions in the later case. The approximation was found adequate since the simulation results of the nonideal response were found to be of close proximity to the experimental results of Fig. 8. The experimental results were obtained using a three bit word for filter topology programmability to select the type of transfer functions. A two words, four bits each, were used for filter parameters programmability where ω_p and Q_p are controlled independently as given in Table 3. Fig. 7 also illustrates a higher order programmability where a fourth order characteristics are shown for a LPF and a Chebychev BPF.

4. CONCLUSION

The novel design described here has resulted in a universal programmable filter that can be digitally controlled to realize almost any practical filter specifications. This is done through the use of CMOS switches controlled by binary codes to program the order of the filter, the filter topology, the filter center frequency and selectivity. The design procedure required developing optimum switching arrangements for the minimum redundancy in components and the least dependence of the filtering function on switching imperfections such as switches stray capacitances and non-zero and non-linear switch-on resistance. Further investigation is being conducted to develop a programmable switched capacitor realization that can allow frequency scaling by changing clock frequency. Work is also in progress for developing an extended bandwidth programmable filter using the composite operational amplifier technique proposed earlier by the author. Such implementation would lead to a very useful monolithic device at moderate cost.

5. REFERENCES

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Filter Type	Y_1	Y_2	Y_3	Y_4	Y_5	Y_6	Y_7	Y_8	Transfer Function
LP	C	C	$Cs\frac{C}{QP}$	C	C	0	0	C	$T_2 = 2\omega p^2/D(s)$
HP	C	C	C	C	0	C	C	$\frac{C}{QP}$	$T_1 = 2S^2/D(s)$
BP	C	C	C	C	0	C	$\frac{C}{QP}$	C	$T_1 = 2(\omega p/Qp)s/D(s)$
N	C	C	C	C	C	0	C	$\frac{C}{QP}$	$T_2 = (s^2+\omega_n^2)/D(s)$
AP	C	C	C	C	C	0	C	$\frac{C}{QP}$	$T1 = (s^2 - \frac{\omega p}{Qp} s + \omega p^2)/D(s)$

Table 1.
The Elements Values for Different Realizations

where $T(s) = N(s)/D(s)$ and $D(s) = s^2 + (\omega p/Qp)s + \omega p^2$

Binary Input	Filter \ Switch	S_1	S_2	S_3	S_4	S_5	S_6	S_7	S_8	S_9	S_{10}	S_{11}	S_{12}	S_{13}	S_{14}	S_{15}	S_{16}	S_{17}	S_{18}
0 0 0	Low Pass	0	1	0	1	1	0	0	1	0	0	1	0	1	0	0	1	0	1
0 0 1	High Pass	1	0	1	0	0	1	0	0	0	1	0	1	0	1	0	1	1	0
0 1 0	Band Pass	1	0	1	0	0	0	1	0	1	0	0	1	0	1	0	1	1	0
0 1 1	Notch	0	1	1	0	0	1	0	0	0	1	0	1	0	1	1	0	0	1
1 0 0	All Pass	0	1	1	0	0	1	0	0	0	1	0	1	0	1	1	0	1	0

Table 2. The Truth Table of the Logic Controlling Elements Relocation Switches

switch control				Qq	Qp
Sa	Sb	Sc	Sd	KΩ	
0	0	0	0	24.0	15
0	0	0	1	22.4	14
0	0	1	0	20.8	13
0	0	1	1	19.2	12
0	1	0	0	17.6	11
0	1	0	1	16.0	10
0	1	1	0	14.4	9
0	1	1	1	12.8	8
1	0	0	0	11.2	7
1	0	0	1	9.6	6
1	0	1	0	8.0	5
1	0	1	1	6.4	4
1	1	0	0	4.8	3
1	1	1	1	3.2	2
1	1	1	0	1.6	1

switch control				C	fp
Sa	Sb	Sc	Sd	nF	KHz
0	0	0	0	5.96	16.693
0	0	0	1	6.34	15.709
0	0	1	0	7.16	13.887
0	0	1	1	7.72	12.875
0	1	0	0	8.42	11.810
0	1	0	1	9.20	10.856
0	1	1	0	10.00	10.043
0	1	1	1	11.00	9.047
1	0	0	0	15.10	6.650
1	0	0	1	17.60	5.637
1	0	1	0	20.80	4.823
1	0	1	1	26.00	3.828
1	1	0	0	35.50	2.805
1	1	0	1	55.00	1.810
1	1	1	0	100.00	.999

Table 3. Binary Words Controlling Qp & fp

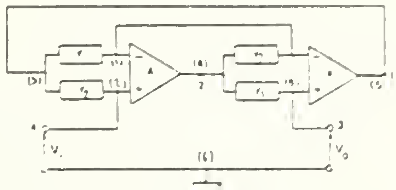


Fig. 1 The Generalized Immittance Converter

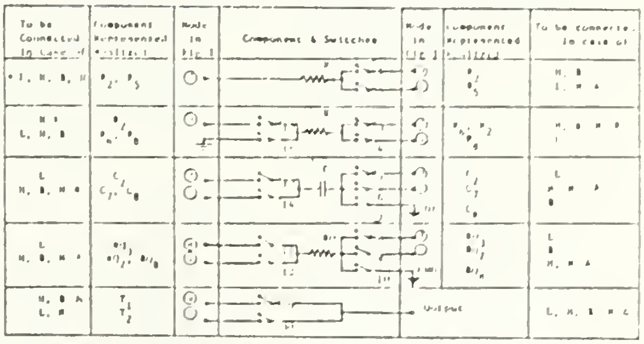


Fig. 2 Elements Relocation Switches for topology programmability

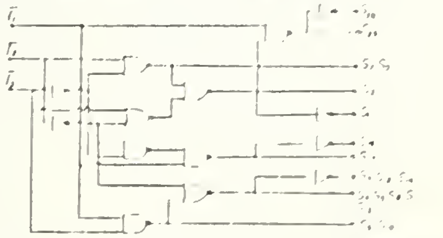


Fig. 3 The logic controlling figure 2 switches

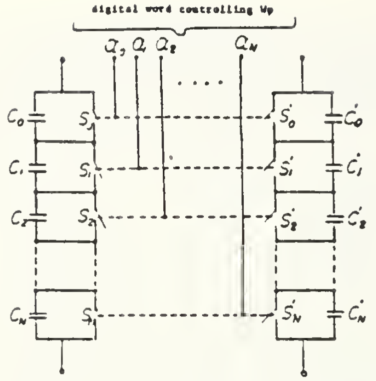
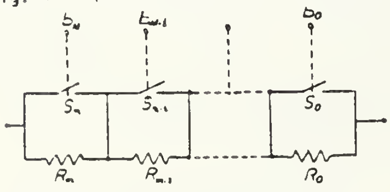


Fig. 4 Capacitors Banks for Frequency Programmability



(For linear Qp control. $R_{j+1}=2R_j$ resulting in $RQp=\sum_{j=0}^{n} R_0 b_j$)

Fig. 5 The Programmable Rq Resistor

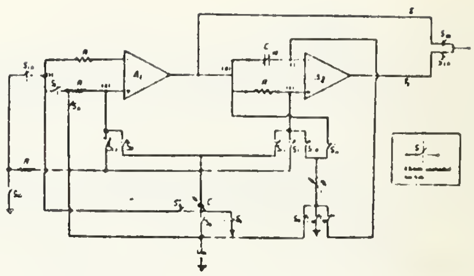


Fig. 6.a The General Second Order GIC Section

ORDER	S_1	S_2	S_3	S_4	S_5	S_6	S_7
2	O	O	O	O	O	O	X
4	X	O	O	O	O	X	O
6	X	X	O	O	X	O	O
8	X	X	X	X	O	O	O

O OPEN
X CLOSED

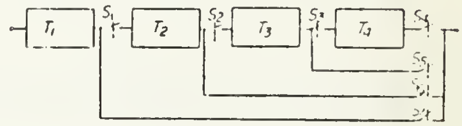


Fig. 6.b Higher Order Programmability of the GIC Sections

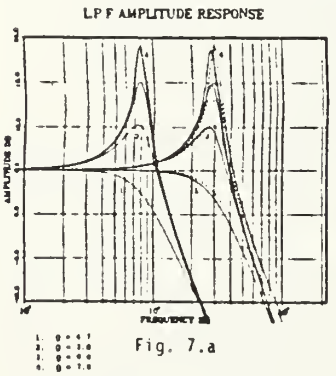


Fig. 7.a

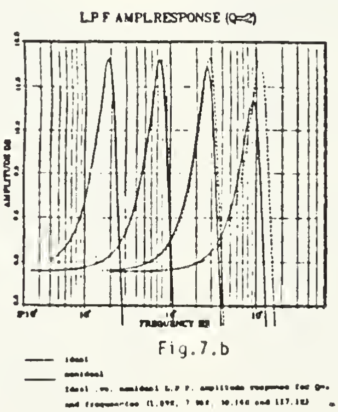


Fig. 7.b

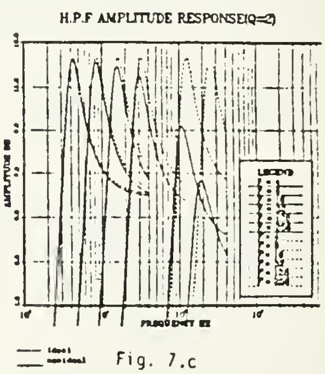


Fig. 7.c

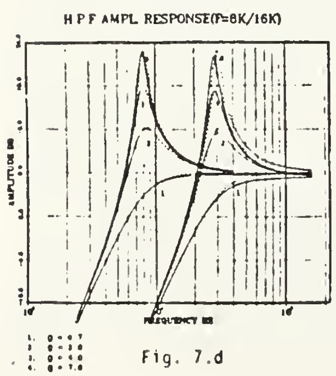


Fig. 7.d

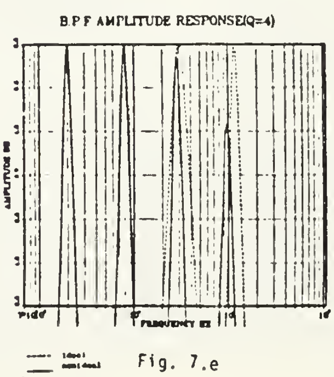


Fig. 7.e

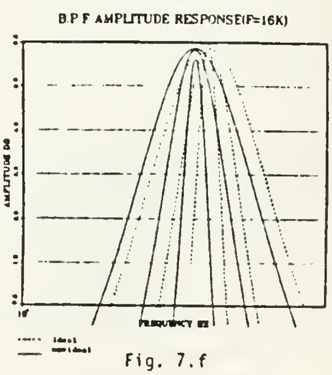


Fig. 7.f

4

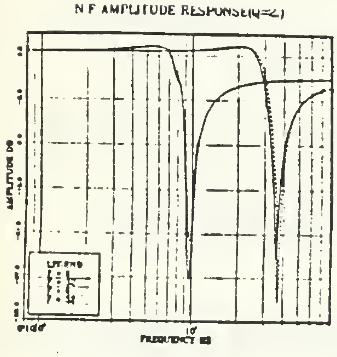


Fig. 7.g

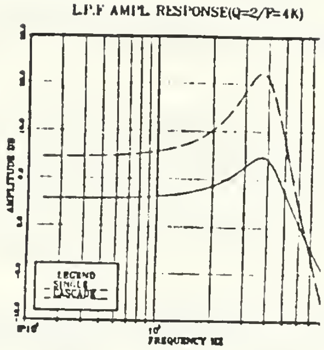


Fig. 7.h

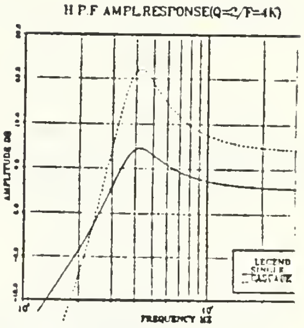


Fig. 7.k

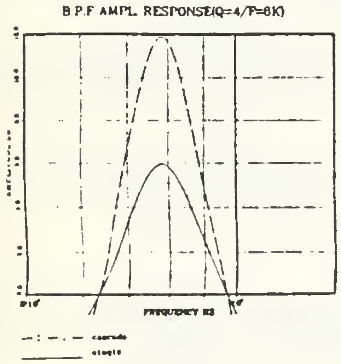


Fig. 7.l

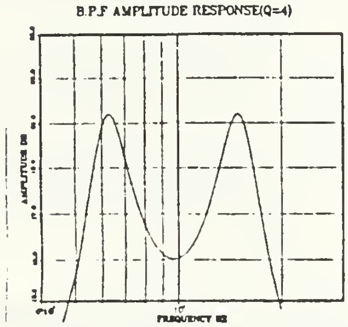


Fig. 7.m

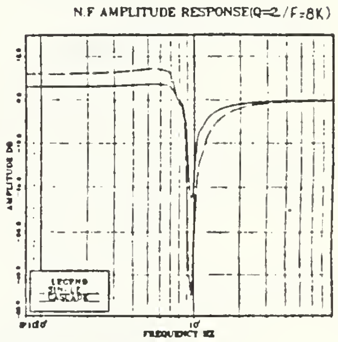


Fig. 7.n

Fig. 7 Computer Simmulation's Results.

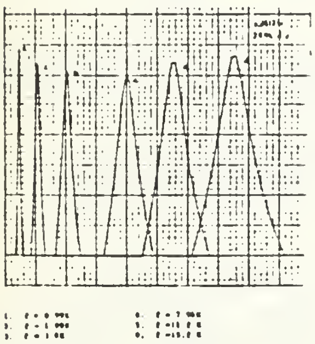


Fig. 8.a

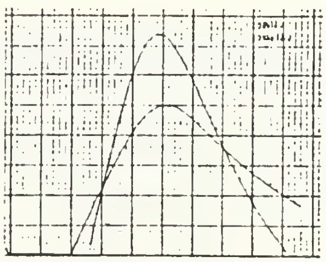


Fig. 8.b

2nd order & 4th order B.P.F

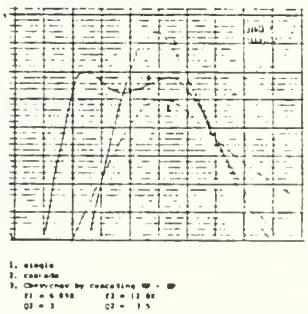


Fig. 8.c

B.P.F Realization for different frequencies

Fig. 8 Experimental Results

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